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TECHNOLOGY UTILIZATION

COMMUNICATIONS TECHNIQUES AND EQUIPMENT

A COMPILATION

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

[redacted]

Foreword

The National Aeronautics and Space Administration has established a Technology Utilization Program for the dissemination of information on technological developments which have potential utility outside the aerospace community. By encouraging multiple application of the results of its research and development, NASA earns for the public an increased return on the investment in aerospace research and development programs.

Compilations are now published in one of nine broad subject groups:

SP-5971: Electronics - Components and Circuitry
SP-5972: Electronics Systems
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SP-5976: Mechanics
SP-5977: Machinery
SP-5978: Fabrication Technology
SP-5979: Mathematics and Information Sciences

When the subject matter of a particular Compilation is more narrowly defined, its title describes the subject matter more specifically. Successive Compilations in each broad category above are identified by an issue number in parentheses: e.g., the (03) in SP-5972(03).

This Compilation is devoted to equipment and techniques in the field of communications. It contains three sections. One section is on telemetry, including articles on radar and antennas; the second section describes techniques and equipment for coding and handling data. The third and final section includes descriptions of amplifiers, receivers, and other communications subsystems.

Additional technical information on items in this Compilation can be requested by circling the appropriate number on the Reader Service Card included in this Compilation.

The latest patent information available at the final preparation of this Compilation is presented on the page following the last article in the text. For those innovations on which NASA has decided not to apply for a patent, a Patent Statement is not included. Potential users of items described herein should consult the cognizant organization for updated patent information at that time.

We appreciate comment by readers and welcome hearing about the relevance and utility of the information in this Compilation.

For sale by the National Technical Information Service, Springfield, Virginia 22161

Contents

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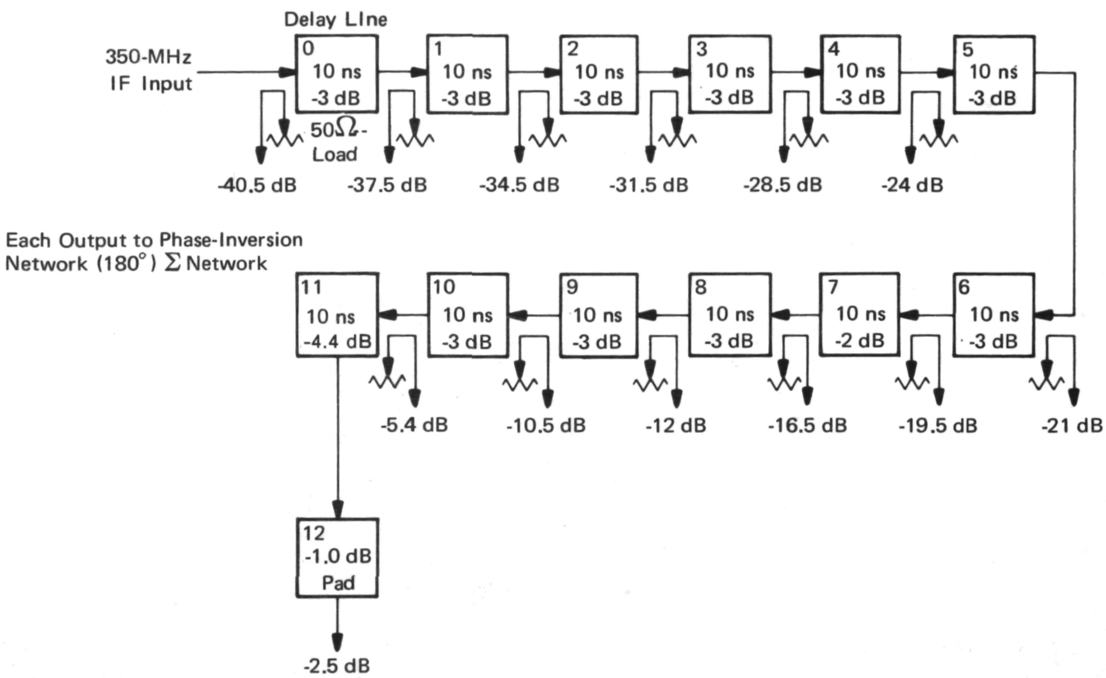
Section 1. Radar, Antennas, and Telemetry

IMPROVED PULSE COMPRESSION FOR A RADAR ALTIMETER



Delay-Line Network

In conventional radar altimeters, the ratio of the peak power of the main lobe signals to the rms power of the side lobe signals is 22.3 dB. A new pulse-compression technique, applied to the signal, increases this ratio to 28.2 dB. This technique requires only moderate modifications to existing techniques.

A 13-bit Barker-Code pulsed signal is compressed to 10 ns, as compared to 200 ns for conventional systems. The transmitted signal is generated with a MECL III logic circuit, at 200 MHz. This signal consists of two simultaneous waveforms, sent from a code generator to a biphase modulator: (1) a 130-ns carrier pulse that controls the duration of the signal pulse and (2) a 13-bit biphase Barker-Code signal that controls the phase of the pulse. The coded transmission is amplified from −10 to +63 dBm in a traveling-wave tube.

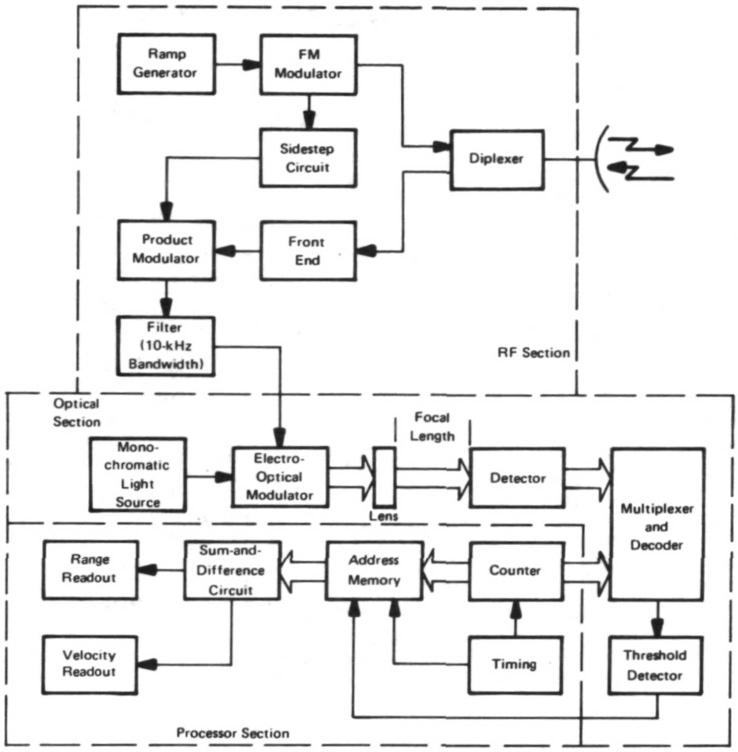
The pulse-compression system consists of 12 coupling elements and an attenuator, all in series with 12 fixed, 10-ns delay-line sections that separate the coupling elements (see figure). The 13 outputs on the line are fed to an inversion network and then are summed coherently in a power divider. The power-divider output contains the compressed pulse.

The improved performance of this system is made possible by several recent developments: (a) the MECL III coherent code generator operating at a 200-MHz clock rate; (b) a balanced RF biphase modulator with switching speeds of 1 to 2 ns; (c) modified traveling-wave tube amplifiers that minimize phase and amplitude modulation; (d) very wideband IF components; and (e) very wideband delay lines, weighting networks, and linear combined networks.

Source: E. A. Dembowski of
General Electric Co.
under contract to
Johnson Space Center
(MSC-14456)

No further documentation is available.

ADAPTIVE DETECTOR FM/CW RADAR



Adaptive Detector FM/CW Radar

The adaptive detector FM/CW radar (see figure) is a lightweight system with a range of 100 km. The continuous-wave (CW) operation permits the efficient use of power while frequency modulation (FM) allows precise range measurement. The linear FM/CW radar transmits a signal of constant amplitude with a frequency which varies linearly over some range. In the detection process, the reflected signal is multiplied with the transmitted signal producing signals for the upsweep and the downsweep at some low frequency within a range given by sweep rate, distance to target, and target velocity relative to the observer. The detection problem consists of finding these narrow-band signals in the allotted range, e.g., 10 kHz, where they are submerged in noise.

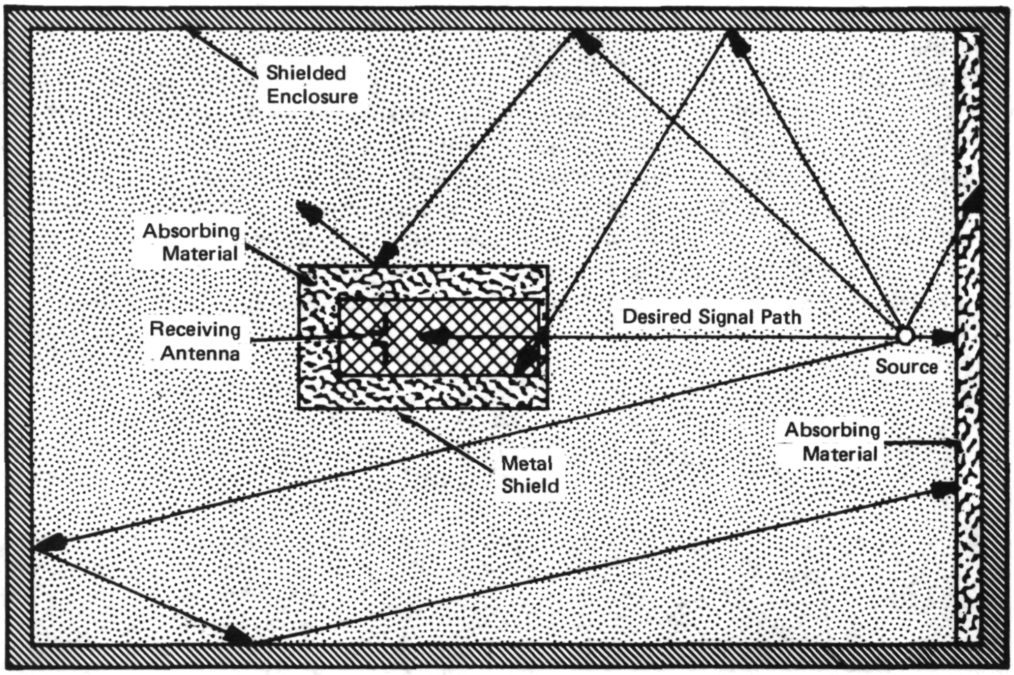
An adaptive optical processor forms a Fourier transform in real time and, by a process of threshold detection, finds the precise value of the upsweep and downsweep frequencies. Doppler and range information are extracted from the sum-and-difference frequencies of the upsweep and the downsweep. The processor consists of an electro-optical modulator which deflects a monochromatic light beam in response to the incident signal, which consists of noise and a sinusoidal voltage at an unknown frequency. The Fourier transform, i.e., the spectrum of the composite input, is produced in the focal plane of a lens placed adjacent to the modulator.

A detector, such as a high-resolution camera or an array of 1000 light-sensitive diodes, senses the amount of incident light. This incident light corresponds to the power spectral density of the composite signal. Each diode represents a band of the spectrum 10 Hz wide. A threshold detector scans the diode array and detects the location of the signal components, which are higher than the uncorrelated noise components in the remaining spectrum. This optical process allows a reduction of the bandwidth from a nominal 10 kHz to 10 Hz, corresponding to a reduction of 30 dB in transmitted power relative to an unprocessed FM/CW radar.

Source: H. Brey and P. E. Geise of
Sperry Rand Corp.
under contract to
Marshall Space Flight Center
(MFS-22234)

Circle 1 on Reader Service Card.

HOODED ANTENNA FOR MEASUREMENT OF ELECTROMAGNETIC RADIATION IN A SHIELDED ENCLOSURE



Hooded Antenna in a Shielded Enclosure

A new method has been developed for measuring 200-MHz to 10-GHz radiation within shielded enclosures. The new approach eliminates the problems of multipath reflections by placing an absorber-lined metal hood around a broadband antenna to shield it from radiation reflected by the enclosure walls. Thus, only the signal traveling the desired path (see figure) reaches the test antenna.

An evaluation of the hooded antenna system for these frequencies shows that measurements made in the shielded enclosure with the hooded test antenna closely correspond to open field measurements. To further demonstrate how multipath reflections in a shielded enclosure are reduced using this technique, the coupling between two antennas has been measured for a spacing of 1 meter in a 2.4- by 2.4- by 6.1-m (8- by 8- by 20-ft) shielded enclosure. Measurements made in the 200-MHz to 10-GHz range using the hooded antenna varied from the open field results by only 2 to 3 dB.

The hooded antenna system can be gain-calibrated by the three antenna method at a 0.3-m (1-ft) spacing, thus allowing correction factors to be used in converting the measured data at the test instrument into field intensity units. This configuration also permits a reasonable degree of control since antenna directivity is a function of the hood aperture-to-wavelength ratio when the antenna is used in this configuration.

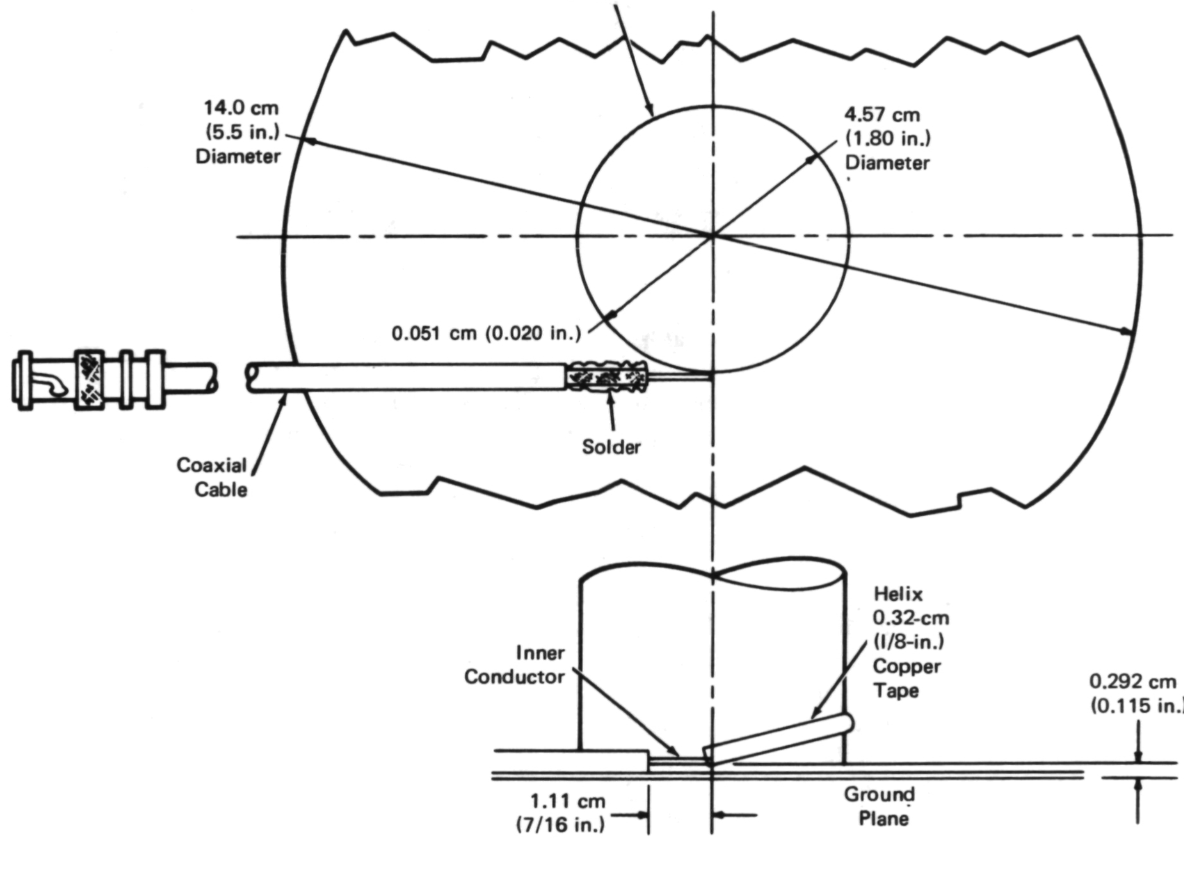
Source: J. C. Toler, W. R. Free, H. W. Denny, J. L. Birchfield, and H. L. Bassett of Georgia Institute of Technology under contract to Marshall Space Flight Center (MFS-21240)

Circle 2 on Reader Service Card.

FREQUENCY-INDEPENDENT FEED FOR HELICAL ANTENNAS

In designing feed systems for helical antennas, it has been a common practice to assume that the helix impedance at the feed point is approximately 140 ohms. Also, it has been assumed that the method of connecting the transmission line to the antenna, as well as dimensions in that area, are relatively unimportant. Measurements have shown, however, that, in order to obtain a low voltage standing wave ratio (VSWR) and broadband matching, the type of connection made between the helical antenna and the transmission line is critical.

A method has been devised for matching a transmission line to a helical antenna over a wide bandwidth, providing several advantages and new features. For example, the impedances are matched outside the coaxial feedline, where more space is available. Moreover, matching sections can be made a part of the helix by printing or painting a helix and matching sections on lightweight fiberglass tube, in a one-step operation. Using this method, the feed system can always be fabricated from standard 50-ohm transmission line. Finally, because the form of the matching section is not unique and because matching sections may be modeled easily using copper tape, experimentation and optimization are simple.



Figue 1. Helical Antenna Feed

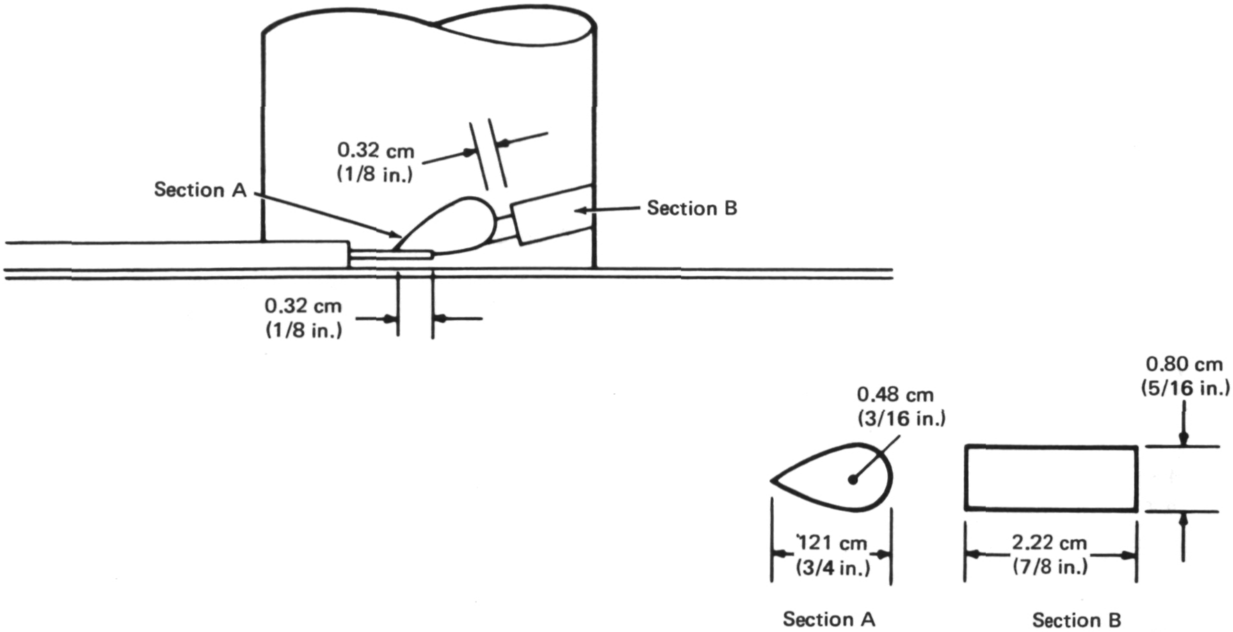


Figure 2. Antenna Matching Sections

The general method is shown in Figure 1. An eight-turn helix is made of copper tape on a fiberglass tube. This helical array then is located at the center of a circular metal ground plane. The feedline is UT 141A semi-rigid, 50-ohm coaxial cable, which is soldered to the ground plane. The inner conductor is bared for approximately 1.11 cm (7/16 in.) and determines the distance between the input end of the helix and the coaxial connector. The coaxial cable itself is positioned in such a way as to be tangent to the circular base of the fiberglass tube.

With the geometry described, the helix input impedance is matched by placing two pieces of copper tape between the coaxial cable and the helix, as shown in Figure 2. Section A of the matching set has a tear-drop form, while section B has a rectangular shape. The matching procedure consists first of positioning section A so that it overlaps the inner coaxial conductor approximately 0.32 cm (1/8 in.) and is lined up horizontally with the conductor (see Figure 2). Section B is spaced about 0.32 cm (1/8 in.) further along the first turn of the helix parallel to the tape. Measurements of the finished antenna show the VSWR to be about 1.3 to 1 over the required 1760- to 2300-MHz band.

Source: Borislav Popovich
Goddard Space Flight Center
(GSC-11027)

No further documentation is available.

SMALL, RECTANGULAR, HORN/TURNSTILE, UHF ANTENNA

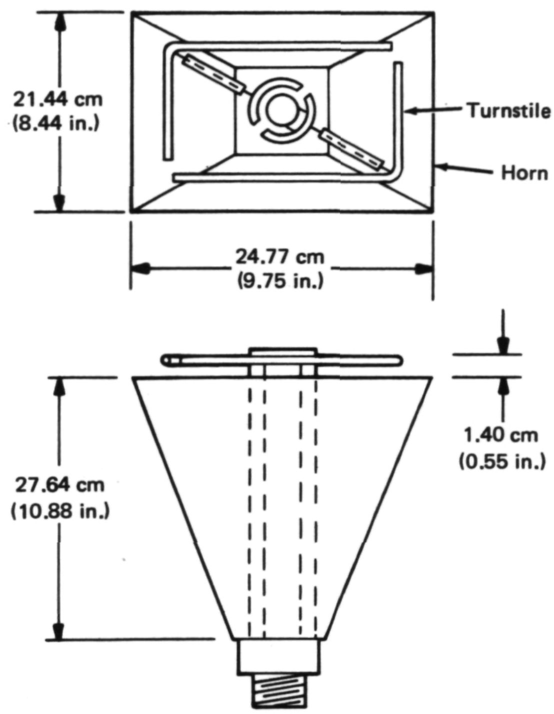


Figure 1. Improved Horn/Turnstile Antenna

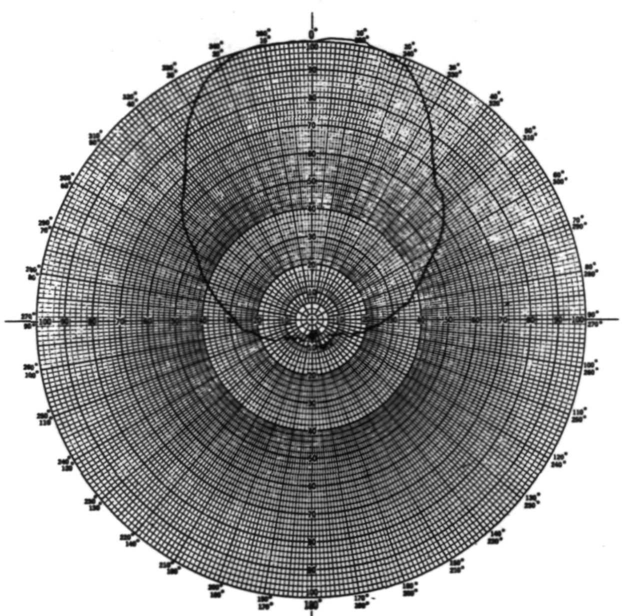


Figure 2. Radiation Pattern: Frequency of 1777 MHz With Linear Source

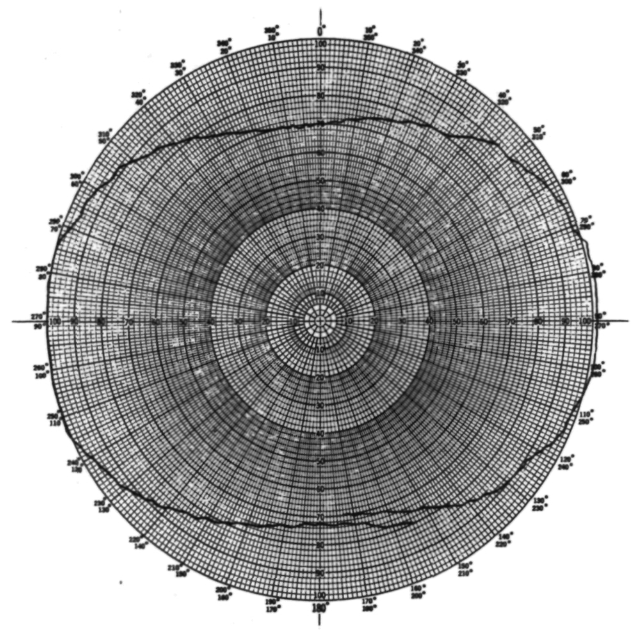


Figure 3. Radiation Pattern: Frequency of 1777 MHz With Source Linear on Axis

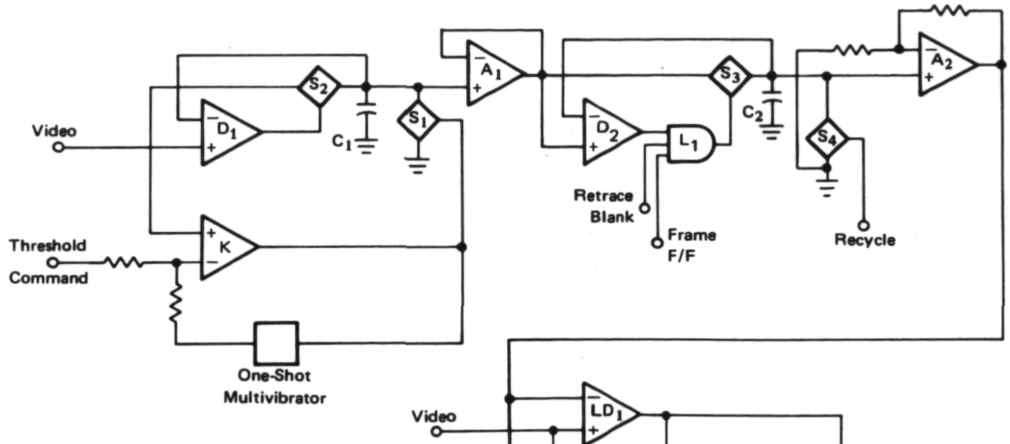
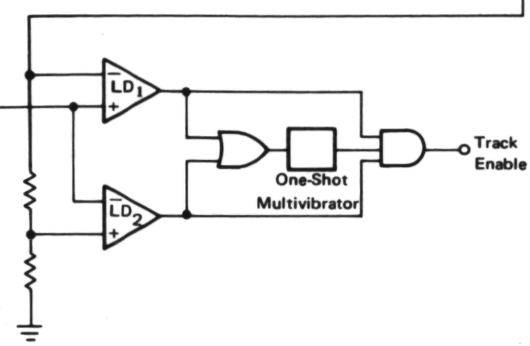
A small efficient antenna was needed for the Viking space vehicle, to provide satisfactory coverage for the UHF communications link during entry, separation, and prelanding operations. Such an antenna might be used also as a feeder for a parabolic antenna. A small, rectangular, horn/turnstile, narrow-band, UHF antenna (see Figure 1) was designed to operate at the transmission frequency during lander descent.

The unique feature of the antenna is the turnstile-feed configuration in combination with a horn-type cavity. The radiation characteristics (for example, azimuth cut pattern and axis pattern circularity) of the prototype antenna are comparable to those of conventional turnstile/cup antennas (see Figures 2 and 3). The dimensions of the rectangular horn are approximately 21 by 25 cm versus an inside diameter of 55 cm for the turnstile/cup antenna. The new design is thus considerably smaller, an advantage from the standpoint of weight and volume.

Source: T. G. Gavrillis of
Martin Marietta Corp.
under contract to
Langley Research Center
(LAR-10973)

No further documentation is available.

REMOTELY CONTROLLED STAR TRACKER WITH THRESHOLD MAGNITUDE DISCRIMINATION

Star-Tracker Threshold Discrimination Circuit

Conventional star trackers are incapable of discriminating between stellar targets according to their relative brightness. The discriminator circuit shown in the figure, however, permits a desired stellar magnitude to be selected and a command to be issued remotely, establishing a brightness threshold. The star tracker then selects the brightest star below this threshold. Selection to within 0.3 magnitude, over a dynamic range of 5 magnitudes, is possible.

The star-tracker system is used as a position sensor, supplying error signals to an active servo system. The system uses a periodically scanned image dissector tube (IDT) and employs an internal target search mode and a track mode. The tracker system produces error signals only when the track mode is enabled. Referring to the figure, a target-generated video signal is compared to a remotely controlled, preset reference threshold in the comparator, K. If the video signal is below the desired level, switch S_1 is opened and capacitor C_1 is permitted to charge to the peak signal level through an ideal diode circuit comprised of operational amplifier D_1 and switch S_2. If the signal is above the desired level, C_1 is shorted to ground and discharged.

Amplifier A_1 acts as a high-impedance input buffer amplifier for the down-stream circuitry. At the end of the first scan across the IDT, switch S_3 is closed, permitting the peak pulse on C_1 to be stored by capacitor C_2. Switch S_4 is normally open unless the target is lost or a new threshold is commanded. The circuit comprised of D_2 and S_3 functions as a second ideal diode.

Amplifier A_2 scales the peak signal stored on C_2 by a factor of 1.2. The output from A_2 is fed to two level discriminators LD_1 and LD_2. LD_2 is set to operate at a level which is 0.8 times the input to A_2. If the video signal into LD_1 and LD_2 on the second and subsequent scan passes has a level between 0.8 and 1.2 times the input to A_2, the logic that follows LD_1 and LD_2 will enable the track mode. This causes an error signal to be generated by the tracker system.

If the video signal on the second scan is less than 0.8, or greater than 1.2 times the signal into A_2, switch S_4 is closed, capacitor C_2 is reset, and the process starts over until two sequential scans occur that verify proper target amplitude. This system has been used to produce servo error signals for fine tracking (to within ±1 arc-minute) on a balloon-borne pointing system. It can be used on any system that is required to track stellar targets or other point sources of light where discrimination between sources is necessary.

Source: R. Cleavinger of
Ball Brothers Research Corp.
under contract to
Johnson Space Center
(MSC-14312)

No further documentation is available.

COMPACT TEMPERATURE SENSITIVE TELEMETRY TRANSMITTER

The layers of air along the surface of a model tested in a wind-tunnel are heated by aerodynamic flow, and on curving around the base of the model, transfer heat to its rear. Figure 1 shows the circuit for a very compact temperature sensitive transmitter. When it is inserted behind the model the resultant temperature changes can be remotely monitored, and the heat transfer rate can be calculated. The circuit is a short-range frequency-modulated (FM) transmitter that changes frequency as the temperature causes the thermistor to change resistance value.

The transmitter is a modified Colpitts oscillator. Miniature chip construction, whereby a thin-film thermistor is deposited on a 0.001-in. (0.0025-cm) nickel substrate, is used to meet the requirements of extreme sensitivity and fast response. This thermistor, when used in conjunction with one or two varactor diodes, causes the oscillator frequency to change in proportion to changes in temperature.

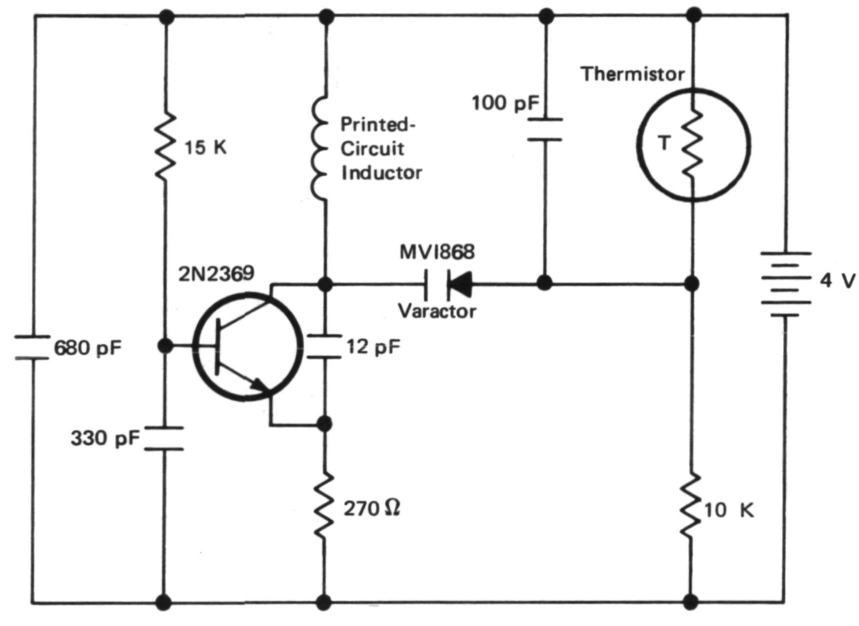


Figure 1. Compact Telemetry Transmitter: Normal Operating Frequency ≅ 115 MHz

$$\text{Sensitivity} = \frac{\Delta\,\text{Frequency}}{\Delta\text{Temperature}} \cong 25\,\frac{\text{kHz}}{°\text{F}}$$

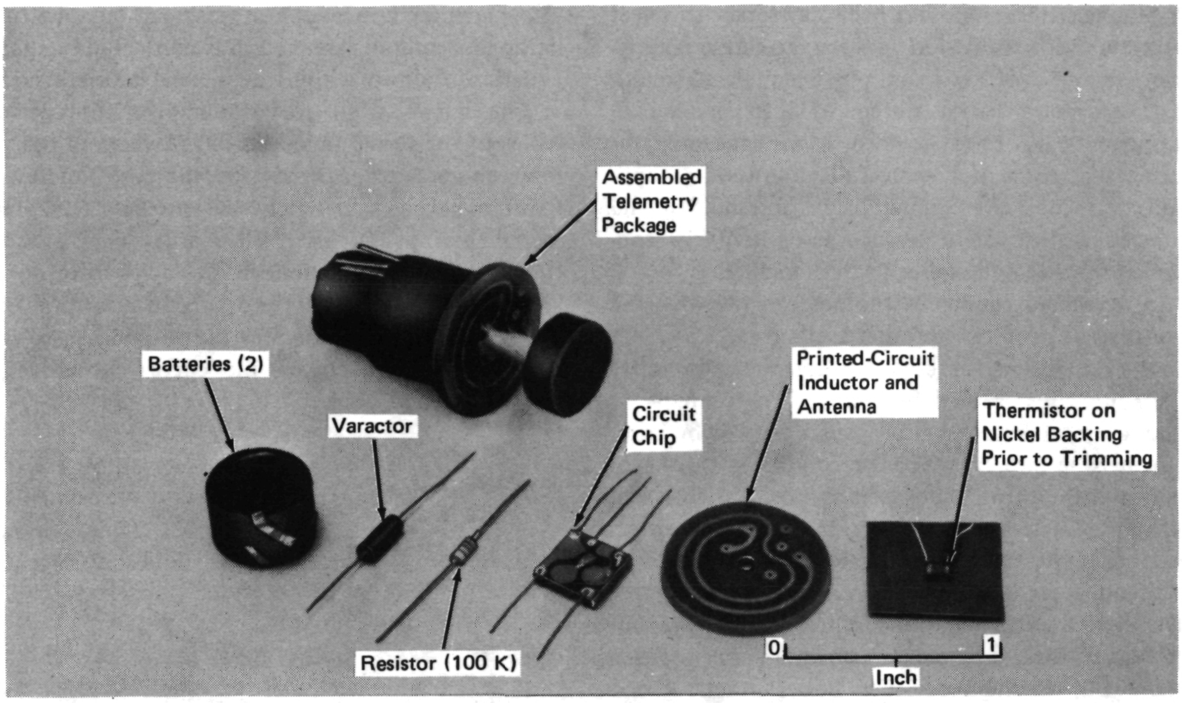


Figure 2. Assembled Telemetry Package and Components (100-pF Capacitor Not Shown)

The small size and rugged construction of the telemetry package (Figure 2) make heat-transfer measurements possible on small models. The transmitter is stable, responds to fast temperature changes, and is capable of operation in extreme environments. The frequency-versus-temperature response is linear within typical tuner (receiver) bandwidths. An alternative oscillator circuit provides a greater frequency change in response to an equal temperature change. The only modifications to the circuit of Figure 1 consist of replacing the varactor with one of a different type and replacing the 100-pF capacitor with a second varactor.

There are many applications for a small, simple (inexpensive) telemeter for remote temperature monitoring. Potential users may be as widely divergent as the petrochemical industry, laboratories, and hospitals.

Source: Royal G. Harrison, Jr., of
NASA Pasadena Office
(NPO-10649)

Circle 3 on Reader Service Card.

OPTIMUM POWER DIVISION FOR SPACE TELEMETRY

A graphical technique has been developed to select the optimum power division between the carrier and the subcarriers of a narrow-band phase-modulated multi-subcarrier communication system. Modulation-loss contour curves have been graphed as design tools for selecting the modulation indices. Two criteria of optimization are considered: simultaneous thresholding and tolerance insensitivity. Tolerance insensitivity is considered as weighted simultaneous thresholding.

The technique eliminates the need for drawing the curves that are required by other methods and provides a graphical solution for both the tolerance-insensitive and simultaneous-thresholding cases. The technique is applicable to n sinusoidal or square-wave subcarriers (where n is an integer) which are considered three subcarriers at a time: two subcarriers as direct variables and the third as a parameter, while all other subcarriers remain constant. The modulation-loss equations derived depend on a parameter K that takes on one of three values: $K = 1$ when no other subcarriers are present; $K = J_0^2(m)$ when sinusoidal subcarriers are present, where $J_0(m)$ is the zero-order Bessel function; and $K = \cos^2(b)$ when square-wave subcarriers are present. The parameters m and b are angular variables. Modulation-loss contours are plotted from the modulation-loss equations defined within the allowable tolerance band.

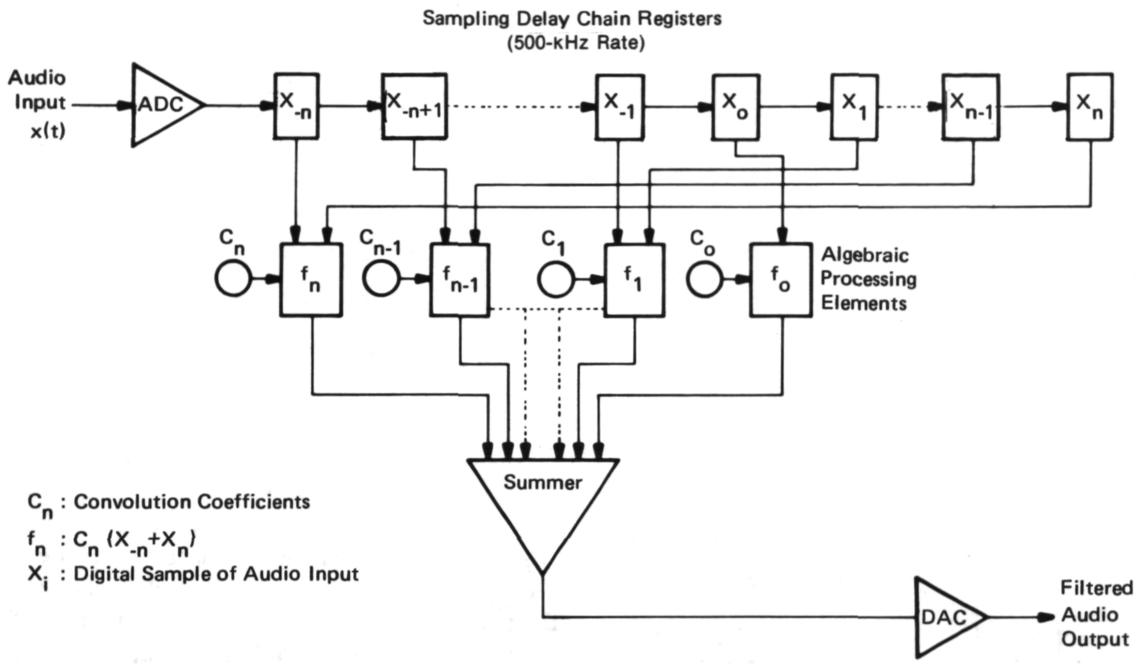
The curves were drawn using computer-generated tables of the Bessel functions of m (where m has values between zero and 2.0 rad) and the $\cos^2$ function of b (with b having values between zero and 1.57 rad, in increments of 0.01 rad). The curves were plotted by rounding off the functions in the tables to three places and then were checked for accuracy to within 0.1 dB. The relevant report contains the formulas used in the derivation of the curves, the curves themselves, and samples of curve usage.

Source: I. Kadar of
Grumman Aerospace Corp.
under contract to
Johnson Space Center
(MSC-12516)

Circle 4 on Reader Service Card.

Section 2. Coding and Data Control

CONVOLUTED AUDIO FILTER



Voice-Convoluted Audio Filter

A digital audio filter has been designed that is based on a mathematical technique for removing random errors from digital statistical data. The filter electronically carries out a moving-average method that uses convoluted integers.

In the moving-average technique, a fixed number of data in a serial sample are averaged. Then a datum at one end of the series is dropped, one is added at the other end, and a new average is taken. The moving average is a special case of a convolution operation, and the set of averages is called the set of convoluted data. The general expression is as follows:

$$Y(j) = \frac{1}{\text{NORM}} \sum_{i=-m}^{+m} C(i)\ Y(j+i)$$

Y(j) is a convoluted datum; the Y(j+i) are the data being averaged; 2m+1 data are being averaged; and NORM is some normalizing factor (for a simple moving average, it is the number of data points). The C(i) are weighing factors, called the set of convoluting integers (for the simple moving average, they are all equal to 1).

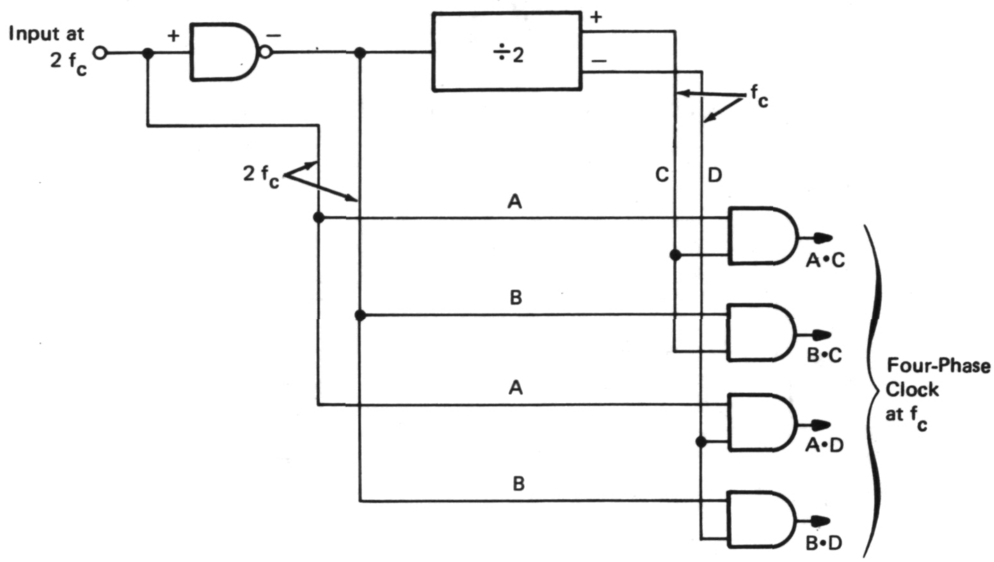
The filter is shown in the figure. The incoming signal is digitized, and a smoothing moving average is taken. However, the C(i)'s and the NORM are calculated from regression or multiparameter optimization theory. The convoluted (i.e., filtered) data are then reconverted to analog form.

The digital filter is a cost-effective solution to problems that are too demanding for analog filters. These can include highly accurate and flexible multiplexing, telephone-line equalization, and possibly the audio programing of computers.

Source: T. R. Edwards and
H. Zeanah
Marshall Space Flight Center
(MFS-22729)

Circle 5 on Reader Service Card.

FREQUENCY-INDEPENDENT, DIGITAL PHASE MODULATION



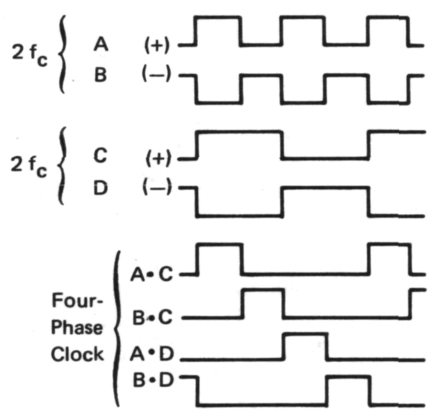


Figure 1. Four-Phase Clock Generator

A new method permits the application of phase modulation to a system clock oscillator in such a way that the percentage modulation is independent of clock frequency. The method is unique because it is digital and completely frequency-independent.

It has several important advantages. Results are completely reproducible; no alinement is necessary. The circuit will work with any clock frequency that is within the speed capabilities of the digital devices used; no tuning is needed. Because it uses digital integrated circuits, it can be made smaller, lighter, and more reliable. For the same reason, susceptibility to environmental change is minimized.

The digital phase modulator combines the generation of multiple phases of the incoming clock signal with digital selection of specific phases in a predetermined sequence. In the example shown, a four-phase clock is used (see Figure 1). The incoming clock signal is twice the desired output frequency. The input, at $2f_c$, feeds a countdown stage. By using the true and complimentary outputs of the f_c and $2f_c$ signals, the four different phases of a clock at f_c are obtained by AND gates. These are present as parallel outputs with timing as shown.

The digital phase modulator is shown in block diagram form in Figure 2. It operates in either of two modes: SEARCH or TRACK. In the SEARCH mode (fixed offset frequency), the reference oscillator is gated off so that inputs A_0 and A_1 are both zero. The outputs from the full adder (Z_0, Z_1) are therefore the same as inputs B_0, B_1.

This output is a 2-bit word derived from the input at two times the desired offset frequency. This word is the control signal for the commutator, causing it to step through the four clock phases once per cycle of the offset frequency. If the commutator is stepped in one direction, the resultant output clock signal will gain extra pulses, having a resultant frequency of f_c+f offset; if stepped the other way, it will lose pulses with a resultant frequency of f_c–f offset. The offset is independent of clock frequency. The search rate is a variable that is constrained only by the inherent speed of the logic elements used in the design.

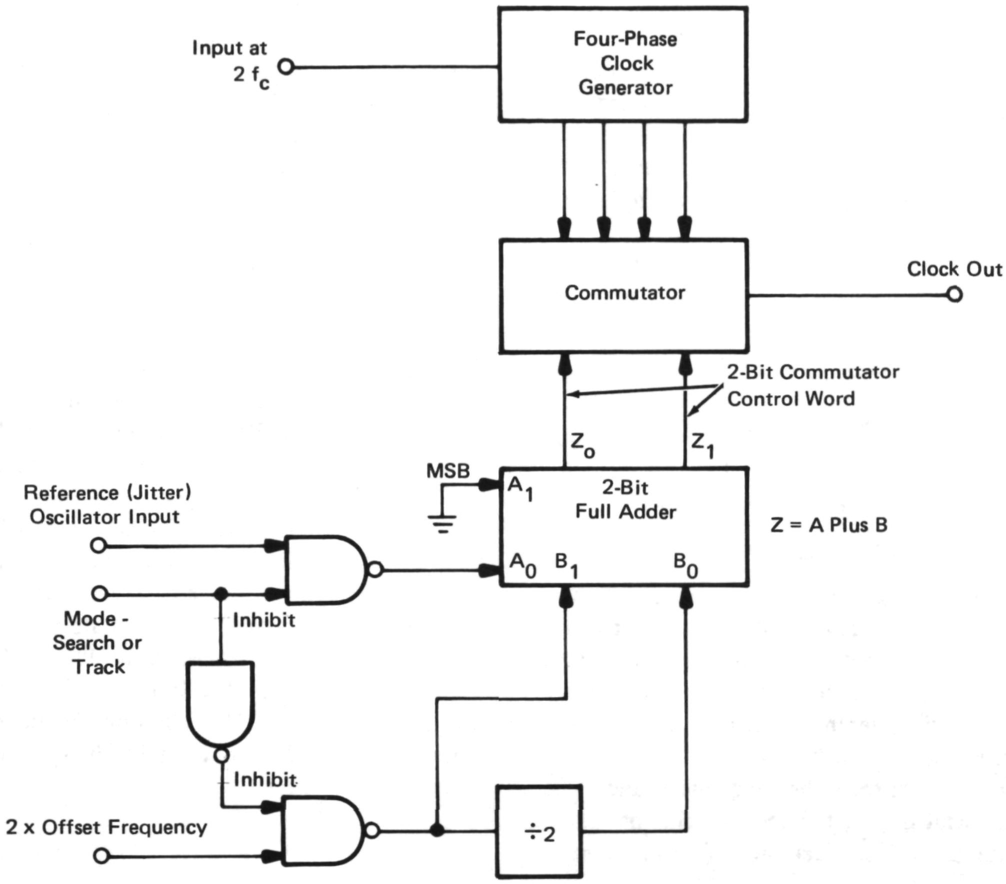


Figure 2. Digital Phase Modulator, Two-Mode (Search/Track) Configuration

In the TRACK mode, a 1/4-bit phase jitter is generated. This is accomplished by inhibiting the offset frequency from the full-adder input B and enabling the reference oscillator to the least-significant bit of input A to the adder. The output is now a 2-bit word that alternates between two adjacent values at the reference oscillator rate. The commutator therefore alternately selects one of two adjacent 1/4-bit phases of the clock at the same rate.

Although the design provides a 1/4-bit (±1/8-bit) phase modulation, this is not the only value possible. Other configurations, such as ±1/4-bit, ±1/16-bit, etc., or even decimal fractions, could be implemented if required for other applications.

Source: J. C. McIlnay of
Westinghouse Electric Corp.
under contract to
Goddard Space Flight Center
(GSC-11251)

Circle 6 on Reader Service Card.

COMMUNICATION SYSTEM CHANNEL SELECTION

The selection of suitable communication system channel frequencies involves many factors, including the consideration of possible interference between channels. A straightforward technique for analyzing the effect of intermodulation (IM) product interference in the selection of channel frequencies has been developed.

The IM product frequency equation has the following form:

$$f_i = \pm C_j f_j \pm C_k f_k \pm C_l f_l \qquad (1)$$

where:

f_i is the channel receiving the interference;

C_j, C_k, and C_l are integer coefficients;

$C_j + C_k + C_l$ is the order of the IM product; and

f_j, f_k, and f_l are channel frequencies.

The amplitudes of IM products decrease rapidly as their order is increased. Consequently IM products of order four and above may be neglected. In accordance with equation 1, all IM products can be evaluated by comparing sums and differences with each other and with channel frequencies.

The number of channels, the frequency range, the signal spectrum widths (f_t), the receiver channel bandwidths (f_r), and the IM product order (m) are used in the frequency selection. The first channel is assigned a frequency at the low end of the allowable frequency range. The next channel is assigned a frequency based solely on the minimum frequency separation (f_{min}) required to prevent adjacent channel interference. Each succeeding channel is assigned a frequency that is separated from the preceding channel frequency as follows:

$$\text{SEPARATION} = f_{min} + (n_i - 1)(mf_t + f_r)/2 \qquad (2)$$

where n_i is the channel rank, and takes on values from one to the number of channels, beginning with the first channel assignment. The IM product order has a value of three. If the signal spectrum widths are not equal, f_t must represent the average spectrum width.

A tabulation of sums and differences which allows the detection of all possible IM products is used in the frequency selection process. The table (see figure) illustrates the technique. Using equation 2, seven channels are assigned frequencies from 11 to 15 kHz. The data rate is such that the transmitted signal-spectrum widths can be ignored. The minimum frequency between channels is 400 Hz and each receiving channel has a bandwidth of 180 Hz.

Channel Freq. (Hz)	11,000	11,400	11,890	12,470	13,140	14,030	14,790
11,000	22,000	400	890	1,470	2,140	3,030	3,790
11,400	22,400	22,800	490	1,070	1,740	2,630	3,390
11,890	22,890	23,290	23,780	580	1,250	2,140	2,900
12,470	23,470	23,870	24,360	24,940	670	1,560	2,320
13,140	24,140	24,540	25,030	25,610	26,280	890	1,650
14,030	25,030	24,430	25,920	26,500	27,170	28,060	760
14,790	25,790	26,190	26,680	27,260	27,930	28,820	29,580

Frequency Selection Table

Excessive IM products occur between the low and high channels because the separation between adjacent channels is near the separation between nonadjacent channels. The two higher channel frequencies are modified, and all but one IM product is eliminated. The two 25,030 entries (circled) indicate the channels involved. This IM product is eliminated by assigning the pertinent channels to functions not operating simultaneously.

The table is constructed as follows (see figure):

a. A matrix is constructed with the selected frequencies entered along the top and down one side.
b. A diagonal is drawn from the top left to the bottom right. Each channel frequency is entered in the diagonal spaces twice.
c. The sums of the channel frequencies are entered below the diagonal.
d. The differences of the channel frequencies are entered above the diagonal.

A comparison of sums with sums and sums with differences gives all third-order IM products. A comparison of sums (diagonal and below) with channel frequencies gives all second-order IM interference.

Source: J. M. Stafford of
IBM Corp.
under contract to
Marshall Space Flight Center
(MFS-21949)

Circle 7 on Reader Service Card.

STATISTICAL PROPERTIES OF FILTERED PSEUDORANDOM DIGITAL SEQUENCES: A REPORT

A report describes the generation of and the properties of pseudorandom digital sequences, and it gives the results of a study of filtered pseudorandom sequences and their statistical properties. The study produced information needed for the design of pseudorandom signal generators. As the output of such generators is produced by filtering long digital sequences, the desirable properties of the signals studied include near-gaussian-amplitude probability density functions and a signal spectral envelope which is approximately that produced by the filter to be used in the generator.

It is shown that sequence generation by the summing of two maximum-length sequences does not always result in a nonskewed amplitude probability density function. Thus it is necessary to study the relation between the digital sequence generator feedback configuration and the sequence statistical structure. The study includes an investigation of the phase distribution of the spectral components of various binary sequences and a direct analysis of the relation between the third moment (skewing indicator) of subsequences of long binary sequences and the sequence characteristic equation. Both approaches yield information useful in evaluating the quality of a pseudorandom sequence.

Source: G. D. Weathers of
Sperry Rand Corp.
under contract to
Marshall Space Flight Center
(MFS-22183)

Circle 8 on Reader Service Card.

RATE DATA ENCODER

An expected range of rate data is determined and set in a new data encoder (Figure 1). Data within this range are encoded by conventional digital techniques. If data fall above or below the expected range (time bias), encoding is stopped and an above or below alert-signal is generated.

Pulses (Figure 2) having frequency, F, where F is to be measured, are applied to the input of the encoder and are delayed by means of the AND gates. The delayed pulses, F_1, are then applied to a flip-flop to produce waveshapes A and B. Pulse train A is fed to a monostable multivibrator (time bias) that produces waveshape C. The period of C is so adjusted that the time interval between the trailing edges of pulses F and C is equal to the reciprocal of the upper bound of the frequency range set in the encoder.

In response to the trailing edge of C, a clock starts producing pulses, G. These clock pulses pass through an AND gate until the occurrence of the next pulse F_1, at which time they are blocked by the voltage level, B. The pulses, G, that pass through the AND gate are counted to give an indication of the frequency of the input pulses, F. The actual encoder value is equal to the count-value algebraic added to the lower bound value. The remainder of the circuitry generates a unique above-or-below alert signal if the frequency of the input pulses exceeds the upper bound or falls below the lower bound established for the encoder operation.

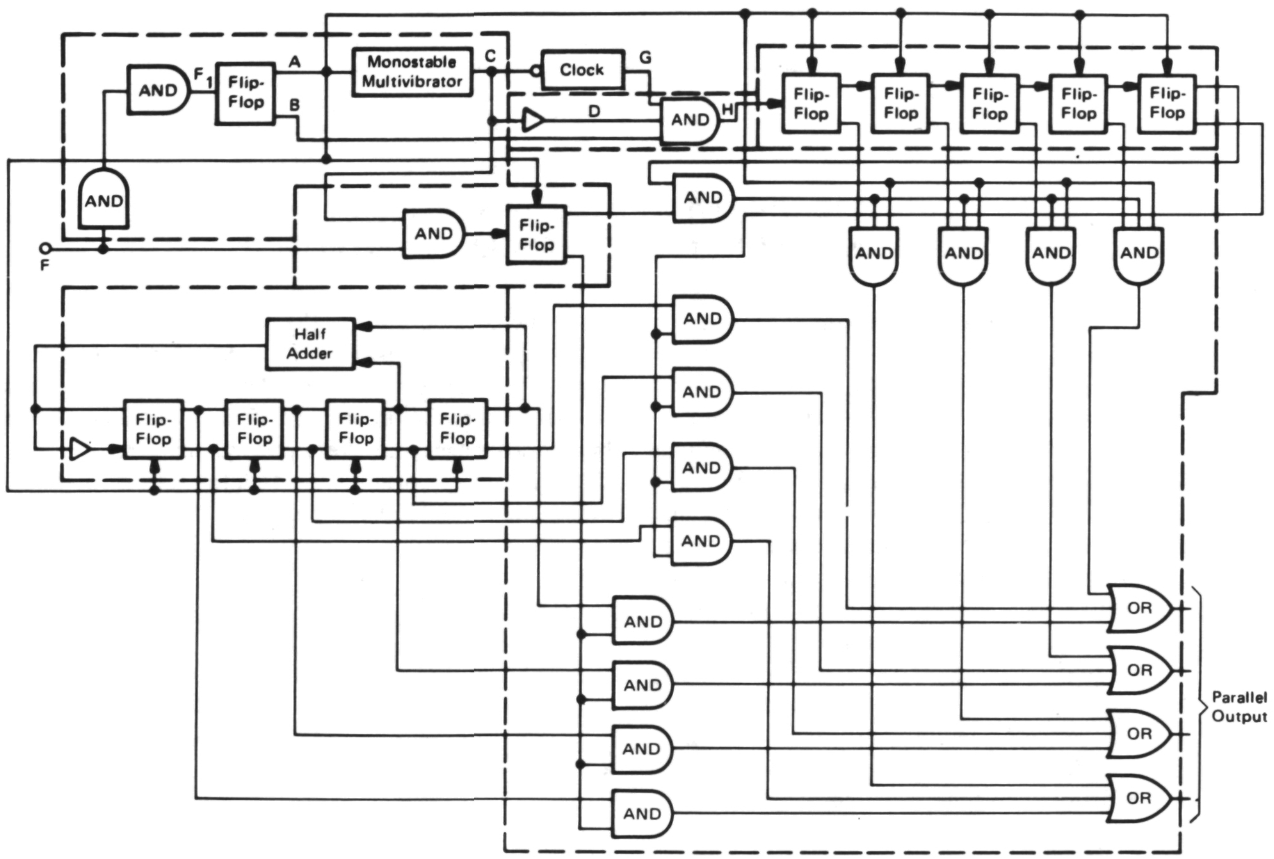


Figure 1. Data Encoder

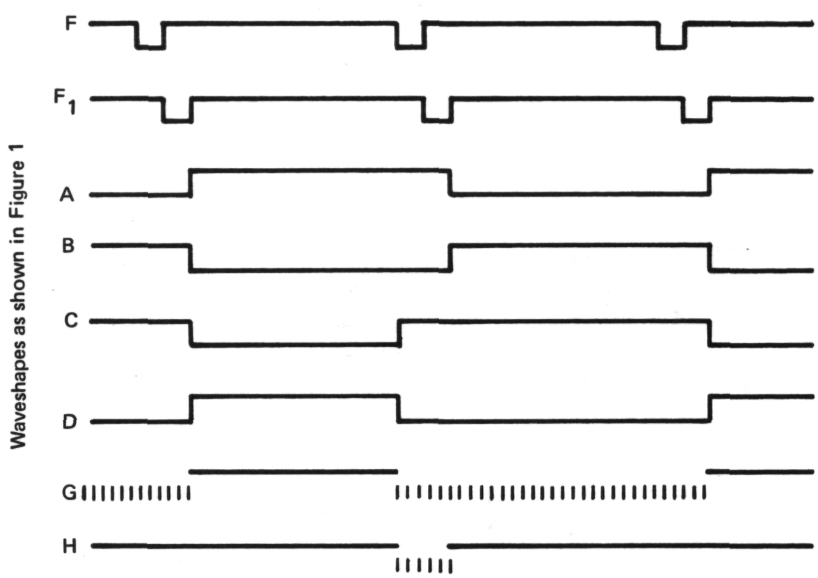


Figure 2. Data Encoder Timing Diagram

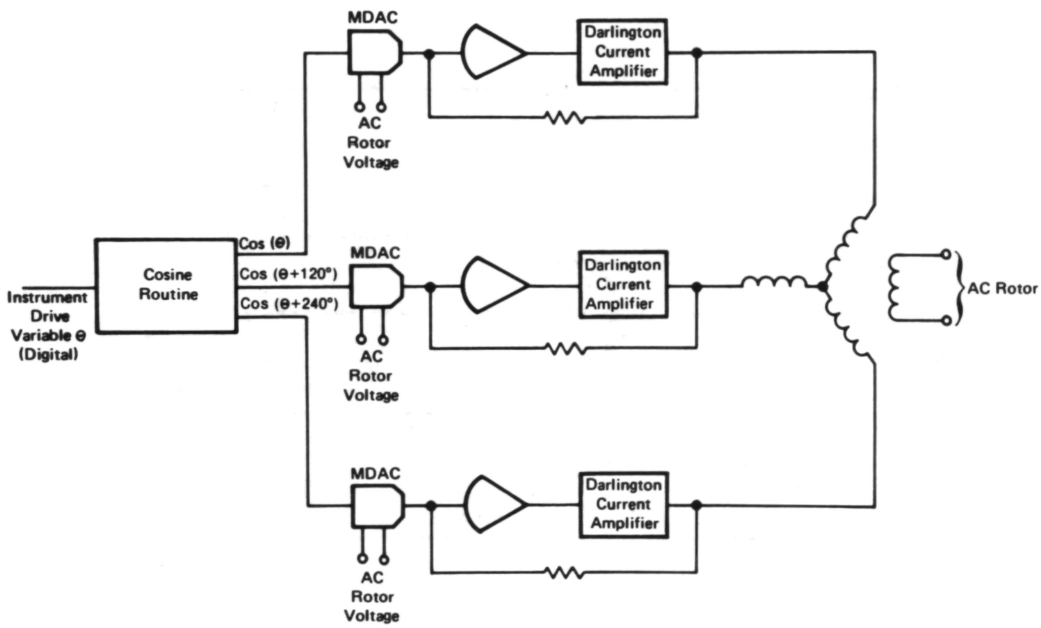
The circuit illustrated encodes only four binary bits of information. Because numbers greater than 15 cannot be encoded in four binary bits, the expected range can be resolved into at most 16 discrete steps. However, this encoder design may be implemented using an increased counter-word length to obtain the resolution required.

This time-bias encoding technique provides a means for reducing the word length (compressing) associated with sampling and coding rate data. The technique can be used to encode rate data such as frequency, pulse rate, revolutions per unit of time, and events per unit of time. This encoder allows the formatting of output data to maintain data accuracy and, where employing a transmission channel, to effect enhancement in transmission efficiency.

Source: W. E. Sivertson, Jr.
Langley Research Center
(LAR-10128)

Circle 9 on Reader Service Card.

DIGITAL-TO-SYNCHRO CONVERTER



Digital-To-Synchro Converter

The converter illustrated transforms digital information into an analog form that is suitable for driving synchro-type instruments. Conventional converters, with gear trains, are inherently limited to angular traverses of less than 360°. There is no limitation on the angular motion of this converter.

The system is based on the fact that the open-circuit phase voltages E_1, E_2, and E_3, which are required to drive the stator of a synchro transformer, are proportional to the transmitter rotor angle θ and the applied ac transmitter/transformer rotor voltage E_r:

$$E_1 = \frac{NE_r}{\sqrt{3}} \cos \theta$$

$$E_2 = \frac{NE_r}{\sqrt{3}} \cos (\theta+120^\circ)$$

$$E_3 = \frac{NE_r}{\sqrt{3}} \cos (\theta+240^\circ)$$

where N equals the synchro transformation ratio.

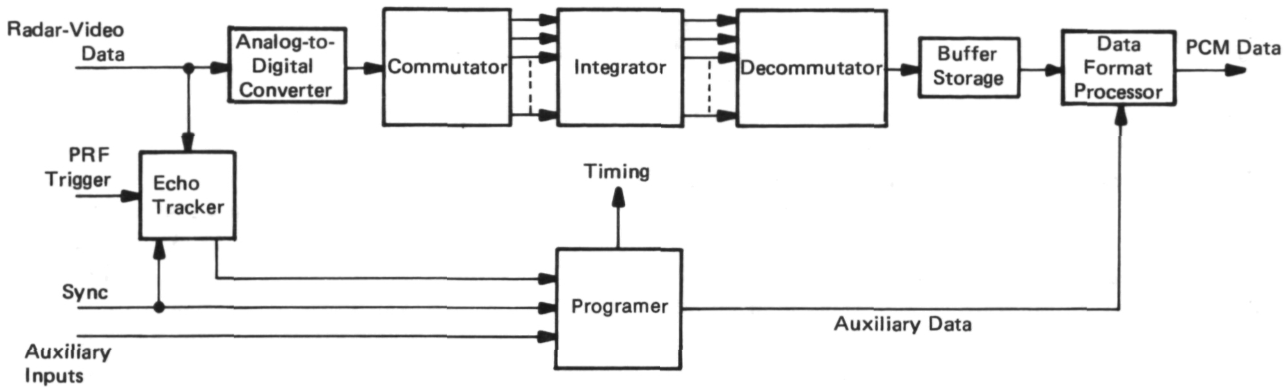
The three cosine terms are generated in the digital computer by a standard cosine subroutine. These values are transmitted to three multiplying digital-to-analog converters (MDAC) where they are transformed into appropriate dc voltages and are modulated by a continuously applied ac rotor voltage. The output of the operational amplifiers associated with the MDAC represents the required synchro transformer drive voltage.

Since synchro transformers are normally low-impedance devices and since operational amplifiers are not normally large-current sources, it is necessary to current-amplify these signals. This is accomplished by a Darlington stage on the output of each operational amplifier. The outputs of the three Darlington stages are the required voltage signals which drive the synchro transformer and which are also fed back to the individual operational amplifiers to keep the Darlington output in compliance with the MDAC output.

Source: Richard D. Murphy of
United Aircraft Corp.
under contract to
Langley Research Center
(LAR-10931)

No further documentation is available.

PREPROCESSOR FOR HIGH-FREQUENCY RADAR-VIDEO DATA: A CONCEPT



Preprocessor Simplified Block Diagram

The preprocessor diagramed in the figure converts radar-video data, obtained at a high sample rate, to a lower rate, compatible with telemetry links and magnetic tape recorders. The processed data retains the accuracy of the original data. The simple, inexpensive solid-state device thus provides real-time radar-acquired information to a home-base terminal.

The radar-video data is sampled, quantized, and converted into a binary format in the analog-to-digital converter indicated in the figure. The output of the converter is a series of pulse-code-modulated (PCM) words, each word corresponding to one sample of video data. The number of video samples taken during each pulse-repetition-frequency (PRF) period equals the number of range-resolution intervals to be interrogated by the radar.

This series of PCM words is time gated, or commutated, into an integrator which provides data storage and adder capabilities. The number of time gates into the integrator equals the number of range-resolution intervals. When the predetermined number of integrations has been achieved, the data is dumped through a decommutator into a storage buffer. The contents of the storage buffer are read out and interlaced with the required auxiliary data by the format processor, at a rate compatible with a telemetry system or magnetic tape recorder.

The preprocessor operates during the dead time between the radar pulses and their corresponding return echo signals, without degrading range resolution and signal dynamic range. It reduces the effective sample rate by integrating an appropriate number of radar echo signals, without degrading azimuth resolution.

Source: S. J. Jarminski and G. C. Covington III of Rockwell International Corp. under contract to Johnson Space Center (MSC-17701)

Circle 10 on Reader Service Card.

PSEUDONOISE TEST SET FOR REAL-TIME AND NON-REAL-TIME SYSTEM EVALUATION USING AN RMS ERROR CRITERION

The pseudonoise test allows the analog-error measurement of communication systems to be performed easily, taking into account the system spikes such as FM "clicks," which are often overlooked by digital methods, and without the excessive distortion often caused by analog approaches. The test set provides both real-time and recorded-test modes and is capable of measuring RMS error, time delay, correlation function, and impulse response.

Filtered pseudorandom data are used with the unit, and an analog error-measuring approach is made to provide near-Gaussian wideband data as a test signal, which exercises the entire designed spectrum of the communication system under test. The unique features of the test set are the following:

a. A continuous variable-delay feature that allows the exact minimization of the RMS error.

b. The utilization of a filter design procedure that allows two filter transfer functions to be matched accurately, and

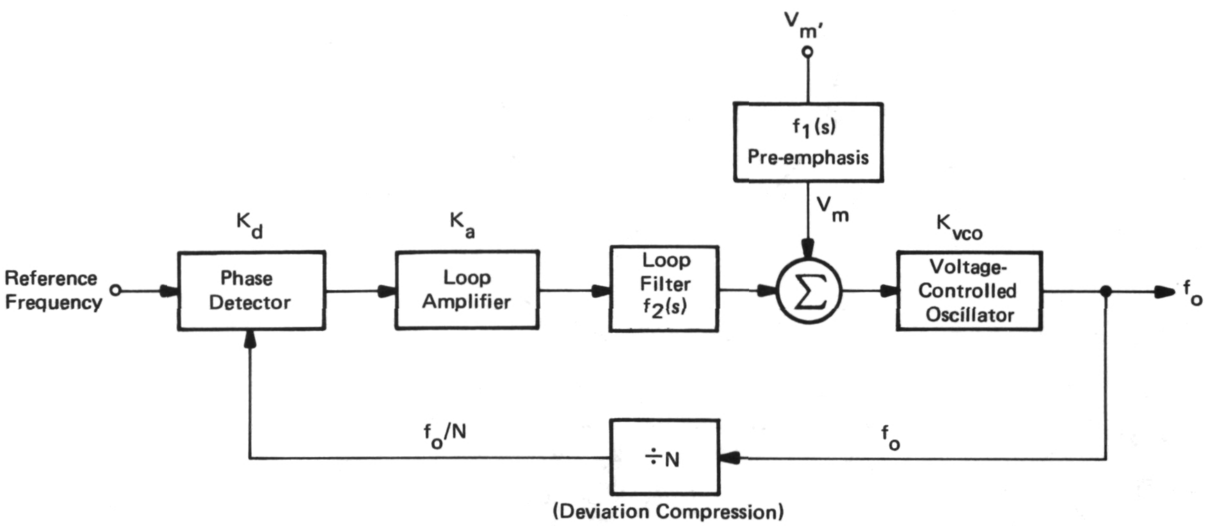
c. The capability for testing recording devices such as tape recorders, using the RMS error-criterion scheme.

Source: G. R. Wallace
Marshall Space Flight Center and
W. E. Salter, G. D. Weathers, and
S. N. Gussow of
Sperry Rand Corp.
under contract to
Marshall Space Flight Center
(MFS-22671)

Circle 11 on Reader Service Card.

Section 3. Communications Subsystems

SYNTHESIZED CLOSED-LOOP FREQUENCY-MODULATOR CARRIER STABILIZER



Carrier Stabilizer Block Diagram

The center frequency drift of an L/C (inductance/capacitance) voltage controlled oscillator (VCO) is unacceptable for many precise modulation applications. A crystal VCO exhibits a stability and center-frequency accuracy that approaches an ideal modulator; however, the maximum frequency deviation is limited. A circuit design technique has been developed for obtaining improved performance from wideband frequency modulation circuits. In the carrier stabilizer diagramed, a feedback and pre-emphasis synthesis technique allows a linear transfer of output frequency to baseband excitation in the frequency region where the baseband excitation and loop bandwidth overlap. The VCO output frequency, f_o, is divided by an integer, N, which is determined by the number of binary dividers in the deviation-compression divider chain.

The frequency of the divider-chain output is f_o/N. This is phase-compared with the reference frequency, and the resulting phase error is amplified, filtered, and used to control the VCO frequency. The net result of this feedback system is that the frequency stability of the reference frequency is transferred to the VCO, within the loop constraints, thus stabilizing the center frequency.

Typical open-loop carrier stability systems are stable to approximately one part in 10^5, referenced to a 1-μs integration time. This closed-loop carrier stability system is stable to one part in 10^8.

Source: G. R. Vaughan and R. C. Ubben of Westinghouse Electric Corp. under contract to NASA Pasadena Office (NPO-10851)

Circle 12 on Reader Service Card.

MONOLITHIC INTEGRATED-CIRCUIT LOG ELEMENT COMPRESSES VIDEO SIGNALS

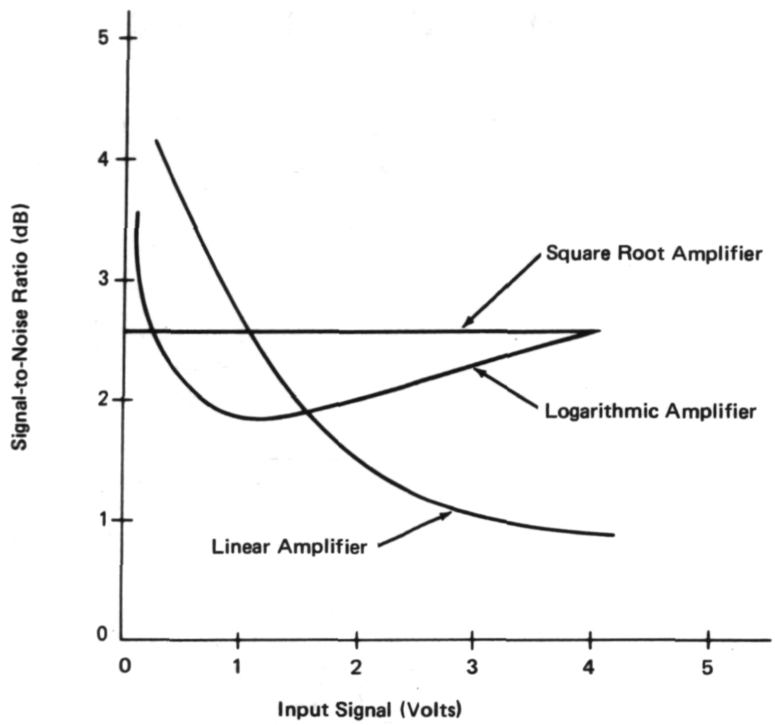


Figure 1. Comparison of Processing Techniques

Video communication systems are frequently limited by shot noise. When the system utilizes a digital quantizer preceeding the modem, degradation in the signal-to-noise ratio may be minimized by nonlinear circuit techniques. When the shot noise of a sensor is expressed as

$$I_n = (2eBI_s)^{1/2}$$

where e = electron charge,
B = system bandwidth, and
I_s = sensor signal,

the resulting signal to noise function is proportional to the square root of the signal. Processing video signals with a linear quantization scheme is inefficient because too many quantization levels exist for small signals and too few exist for large signals. This inefficiency may be eliminated by nonlinearly compressing the signal prior to quantization and processing the signal at the receiver with the inverse function.

The signal can be processed by passing it through a square-root function amplifier. This makes the RMS voltage constant. However, the system efficiency can be further improved for the signal of greatest occurring frequency if log-function processing is used.

In Figure 1, signal-to-noise degradation is plotted for linear, square-root, and logarithmic processing. In the system using the logarithmic amplifier, the signal range is from 0 to 4 volts; but most of the signals are clustered near 1 volt. Thus, it can be seen that for this case, the logarithmic processing will be more efficient than the square-root method.

An integrated circuit with an SN56502 log amplifier and external conditioning circuitry can generate logarithmic transfer functions with wider bandwidths and lower temperature coefficients than are achieved with conventional techniques. The circuit is shown in Figure 2.

This circuit provides a transfer function of the form:

$$V_0 = K_1 \log_{10}(V_i + K_2)$$

where V_0 is the output voltage,
V_i is the input voltage, and
K_1 and K_2 are constants.

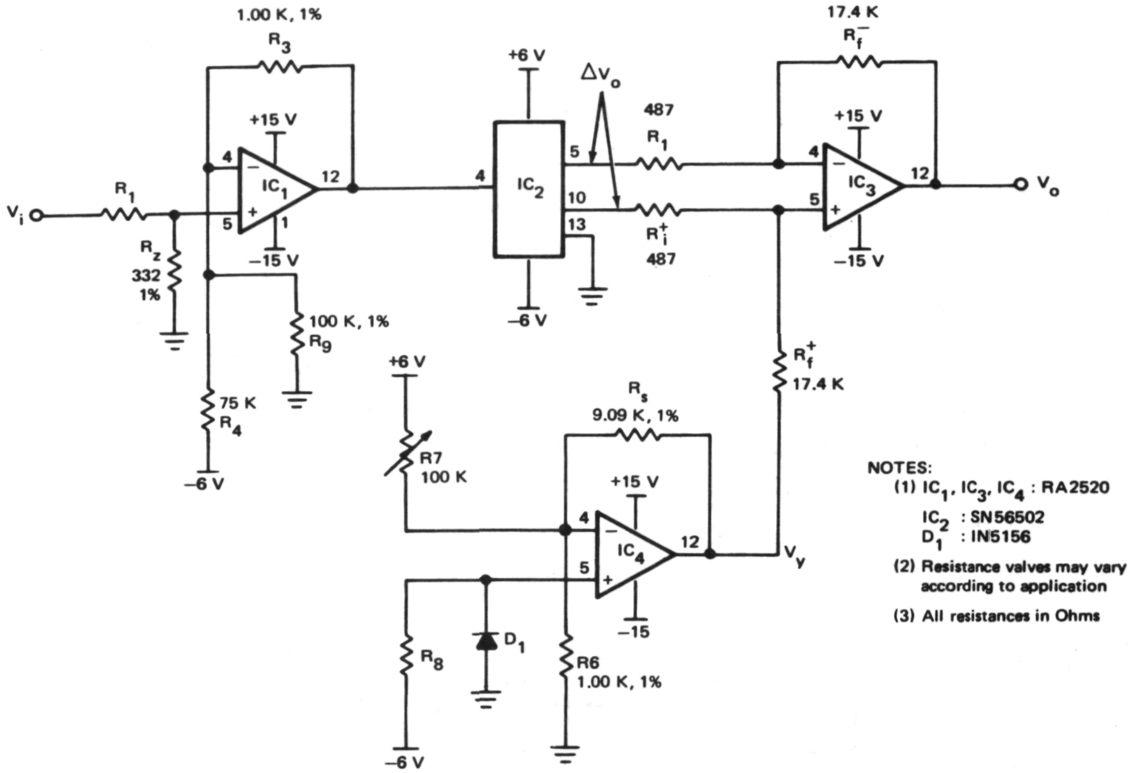


Figure 2. Logarithmic Amplifier Circon

The voltage to be compressed is attenuated at resistor divider network R_1 and R_2. In order to complete the argument of the transfer function (V_i+K_2), the output voltage of IC_1 (V_x) is offset, amplified, and converted to a logarithmic differential voltage, V_0, by IC_2,

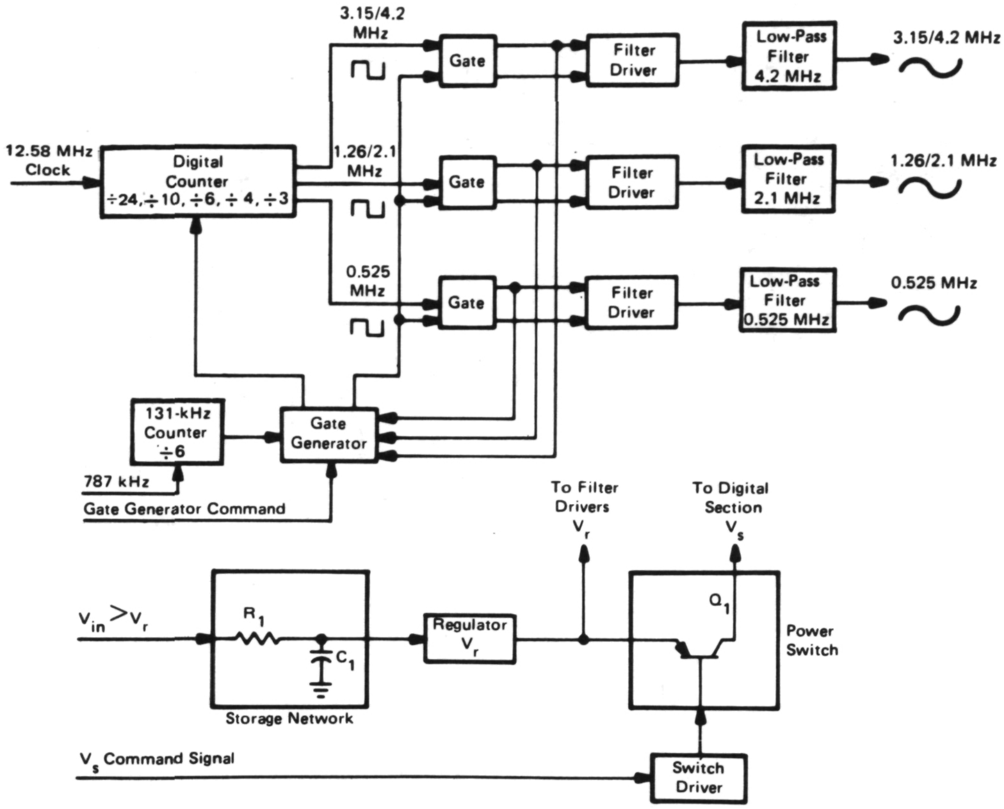
Since the output of the log element of IC_2 is driven by an internal common-base stage with collectors resistively loaded to the +6 volt supply, a common mode voltage of 5.6 volts appears at ΔV_0. In addition to this common mode voltage, an offset appears due to the constant K_2 required at the input to IC_2. The signal ΔV_0 must therefore be inverted, amplified, and level shifted by IC_3.

In order to null out the offset and compensate for the small temperature coefficient of the log element, an offset and temperature control amplifier drives the bottom leg of positive feedback resistor R_f^+. Temperature compensation is provided by the negative temperature coefficient of diode D_1, which is forward biased. Both the forward bias and its variation with temperature are amplified by IC_4. Resistor R_7 allows minor trimming of the output voltage of IC_4, which in effect controls the IC_3 output offset voltage.

Source: John B. Frost of
Hughes Aircraft Co.
under contract to
Goddard Space Flight Center
(GSC-11303)

Circle 13 on Reader Service Card.

FILTERED SQUARE-WAVE MULTIBURST DESIGN APPROACH FOR TELEVISION CAMERA VERTICAL-INTERVAL TESTING



Multiburst Vertical-Interval Test-Signal Generator

The generator illustrated in the figure produces a sinusoidal multiburst waveform that is suitable for use as a vertical-interval test signal (VITS), to monitor the electrical characteristics of the video transmission channel between a remote television camera and a base station. The waveforms produced exhibit low waveform distortion and highly stable frequencies, phases, and amplitudes. The system has low power dissipation, operates over a wide temperature range, and has a high packaging density.

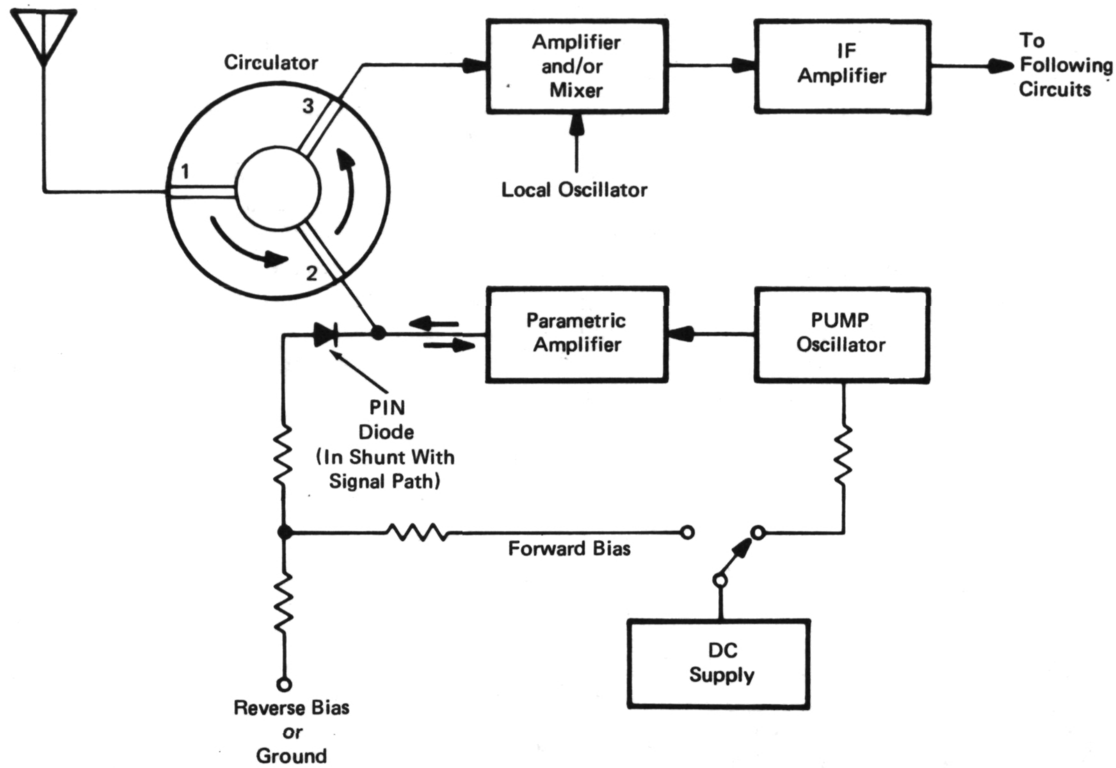
A 12.58-MHz system clock drives a frequency counter in which five counting moduli are used to produce five burst frequencies. Gate generators provide the basic frequency at which the burst frequencies in the multiburst signals are changed. To select the required signal, one of three gates is enabled. The gates provide a two-bit binary control word that commands each of four possible filter-driver stable states. The filter drivers, for which system ground and voltage V_R are used as reference levels, convert the gate outputs to stable, precise voltage levels. Low-pass filters reject the harmonics of the resultant square-wave inputs, while passing the fundamentals as low-distortion sine waves. The three burst outputs can be combined onto a single line by a summing network or an amplifier.

To reduce power dissipation, a power switch turns off the digital section supply voltage except during the burst interval. The voltage regulator provides isolation from camera-supply noise. Low distortion sine-wave bursts in the 0.5- to 4-MHz range are generated by the apparatus.

Source: S. M. Engel of
Westinghouse Electric Corp.
under contract to
Johnson Space Center
(MSC-14329)

Circle 14 on Reader Service Card.

MODE-SWITCHABLE RECEIVER FRONT END



Mode - Switchable Receiver Front End

The figure illustrates a means of maintaining the operation of a radio receiver that is equipped with a parametric amplifier front end, should the amplifier or the high-frequency (PUMP) oscillator degrade or fail. The new feature is a positive intrinsic negative (PIN) diode, in shunt with the signal path at port 2 of the circulator.

In the unbiased mode the diode has little or no effect on the normal operation of the system; in the biased condition the diode appears as an extremely low resistance (a short circuit) across port 2. Since port 2 has been terminated in a nominal characteristic impedance, a short circuit creates an impedance mismatch that results in a reflection of the signal power normally absorbed in the port 2 termination. The signal therefore continues to circulate and is available at port 3 for extraction to the mixer stage. The switching of the bias potential from the oscillator to the PIN diode can be manual, automatic, or by remote command.

Source: Willis S. Campbell
Goddard Space Flight Center
(GSC-10765)

Circle 15 on Reader Service Card.

LOGARITHMIC AMPLIFICATION/DETECTION IN DATA-BUS MODEMS

A data-bus modem operates at fixed-output radio-frequencies between 100 and 450 MHz and processes data that is amplitude modulated on the output RF signal. The modem communicates by way of a data bus with similar modems, provided that all the communicating modems are tuned to the same frequency. Data are transferred between modems at 10-megabits-per-second (Mbps). The receiving data modem must acquire the 10-Mbps amplitude-modulated RF signal and provide a stable recovered data output within a 1.0 μs time period and within an RF signal input dynamic range of 50 dB. The modem consists of a transmitter section, a receiver section, and a logic and control section (Figure 1).

The receiving section of the modem provides the means for acquiring and processing the 10-mbps modulated RF signal. In general this is done by a linear RF amplifier with automatic gain control (AGC). AGC is developed at the output of the amplifier and is fed back to one of the amplifier stages in such a way that the presence of a signal tends to lower the amplifier gain. The time required for this reduction in gain to take place is determined by the constants of the circuit, as well as by the strength of the signal. A single pulse does not appreciably reduce the gain of the amplifier, but long blocks of signals do. Thus, gain is not controlled consistently over the dynamic range of the linear amplifier within the initial time of signal reception.

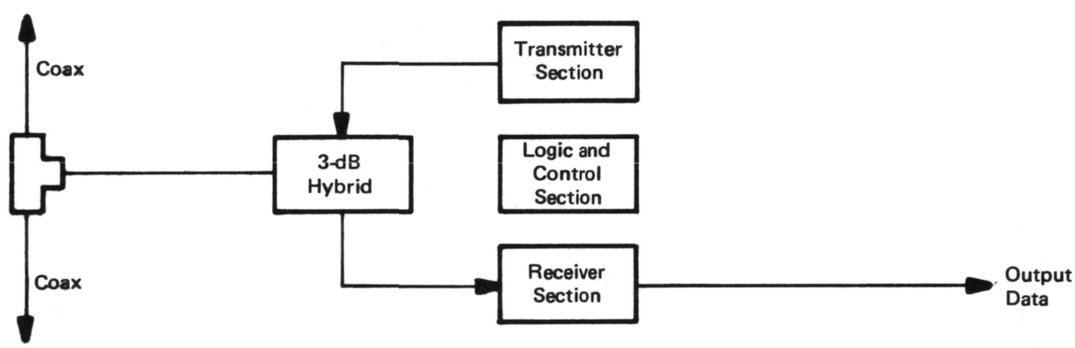


Figure 1. Modem

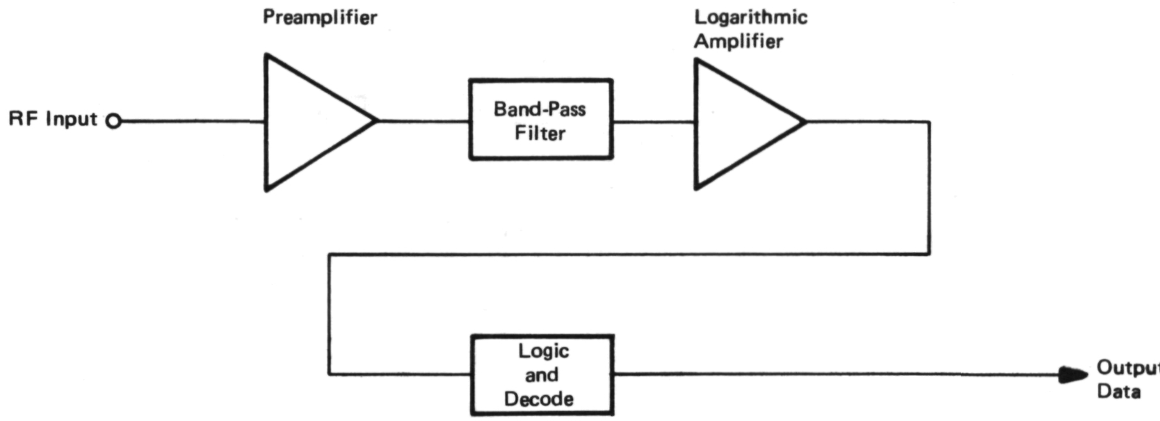


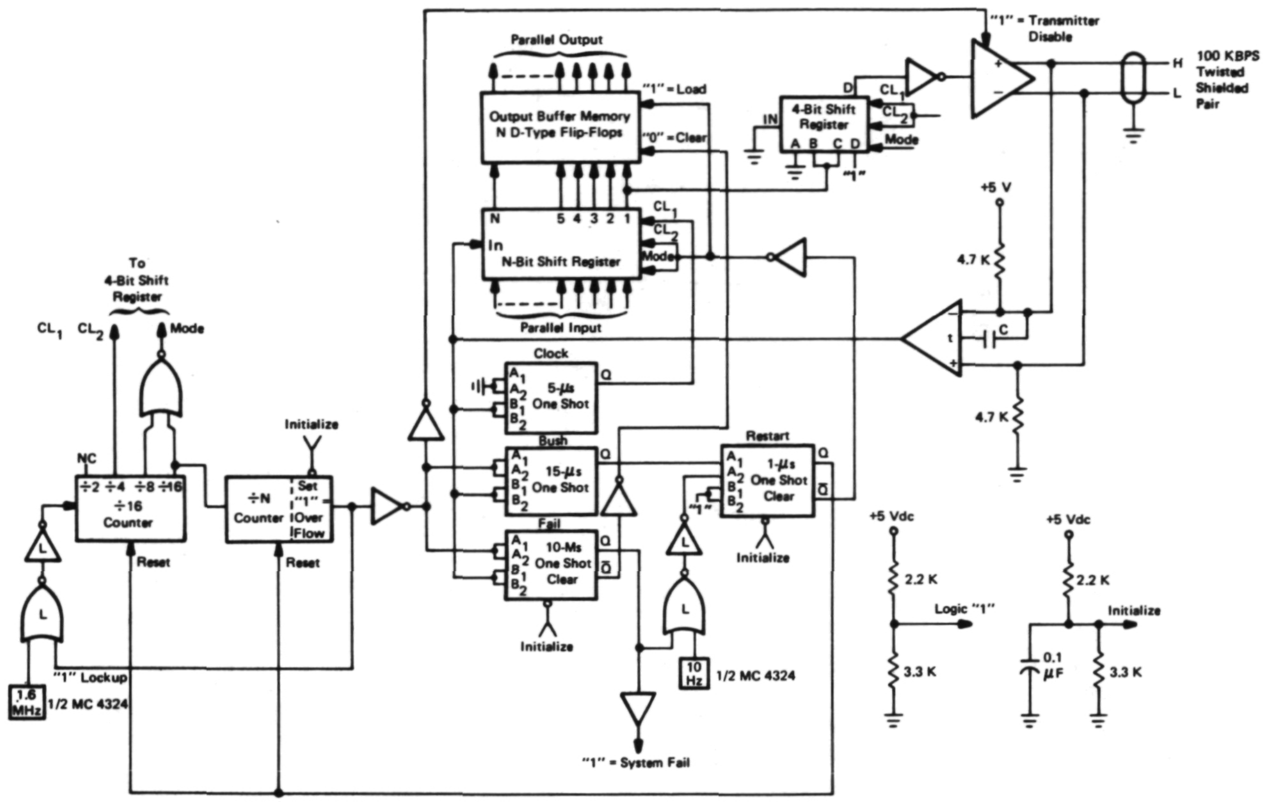
Figure 2. Modem Receiver Section

A logarithmic amplifier, which has gain characteristics such that the output response varies approximately as the logarithm of the input RF signal, will give a relatively constant output level over a large dynamic range without the use of AGC. The amplifier provides consistent gain control by functioning as an amplitude modulation detector in the receiver section of the modem (Figure 2). It recovers the original modulating signal with less than five sync bits lost (the acquisition time is 0.5 μs since each sync bit required 0.1 μs at 10 Mbps), with an input signal dynamic range of -17 dBm maximum to –67 dBm minimum, and without adjustment of the receiver.

Source: L. J. Rettinger and J. B. Whited of IBM Corp. under contract to Marshall Space Flight Center (MFS-22755)

Circle 16 on Reader Service Card.

UNIVERSAL DIGITAL COMMAND/TALKBACK SYSTEM: A CONCEPT



Universal Digital Command/Talkback System

A command/talkback system transfers digital data in two directions over a single hard-wire channel. Two identical units, which are connected by a single twisted-pair shielded line, can transmit serial data in both directions over distances greater than 1 km at the rate of 100 kilobits per second. For an 8-bit system, each unit consumes 1 W at 4.5 to 5.5 Vdc and operates over a temperature range of from –67 to 257° F (–55 to +125° C). The unit can be packaged in a volume of 4 in.^{3} (65.5 cm^3).

Although true simultaneous bidirectional data flow on a single hard-wire channel requires frequency-division multiplexing, virtually simultaneous bidirectional flow is achieved by rapidly changing the direction of data flow. The system provides automatic initialization and continuous reacquisition, and incorporates bit-by-bit clocking and missing-pulse synchronization. A common shift register has automatic clocking for shift control during both transmission and reception. The shift register output consists of data-pulses of ultrastable width.

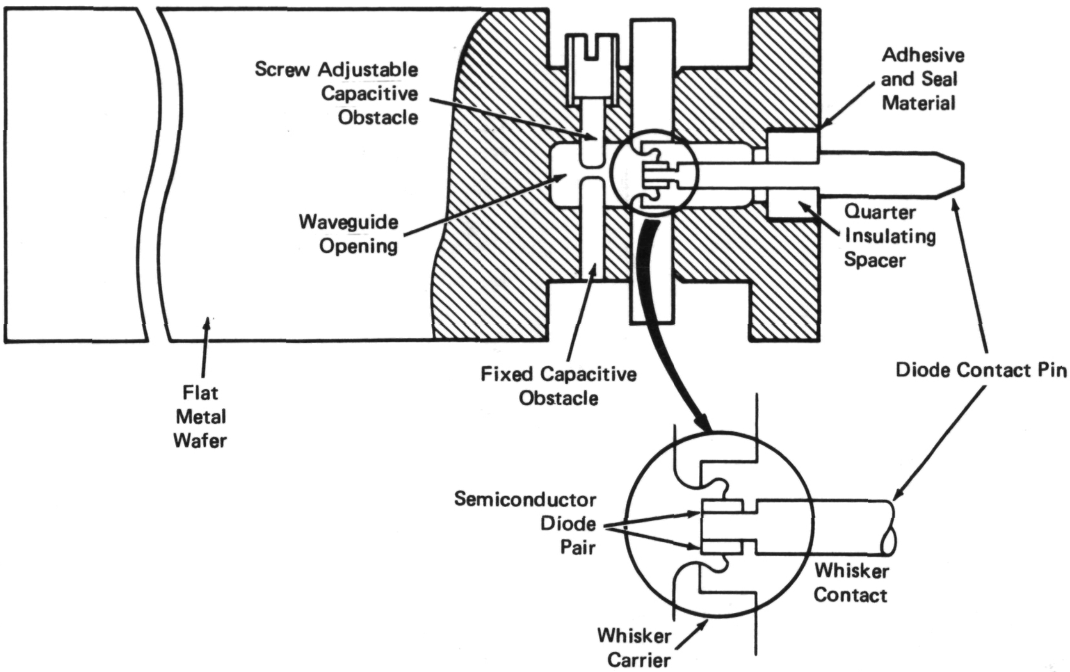
Alternative applications include a party-line configuration, permitting one master unit to address one particular slave unit and receive a reply from it only. In this case each slave is assigned an address, the number of bits required depending on the number of slave units desired. All slave units receive all transmissions from the master unit, but only the slave unit addressed stores the word and returns a reply. When talkback is not required, a master unit could operate in transmit-only mode, talking to several slave units.

A pulse-width modulation scheme is employed, making the system insensitive to data-pulse repetition-rate changes and pulse-width changes in the one-shot multivibrators. Not only may systems using any word length be handled, but also the flexibility in the assignment of bits makes the device suitable for remote-control applications.

Source: R. D. Glover
Johnson Space Center
(MSC-12505)

Circle 17 on Reader Service Card.

VARACTOR DIODE ASSEMBLY



Varactor Diode Assembly

The varactor diode assembly shown in the figure overcomes the parasitic reactances of conventional varactor packages. In this specially constructed assembly very high idler-frequency to signal-frequency ratios are used to obtain low-noise operation over a maximum bandwidth. The idler energy is restricted to the vicinity of the metal wafer holder used for mounting a pair of semiconductor chips. The dimensions of the diode contact wires and the diode junction capacitances are such that resonant frequencies are greater than are possible with conventional diodes.

The enlarged area of the figure shows a pair of semiconductor chips mounted on a contact pin. Contact is established through a pair of low-inductance wire whiskers that are positioned by whisker carriers. An insulating quartz spacer, which is solidly affixed by adhesive and sealer material, positions the contact pin assembly. The two diodes provide a balanced arrangement and geometrically isolate the idler and the fundamental pump energies from the signal circuit.

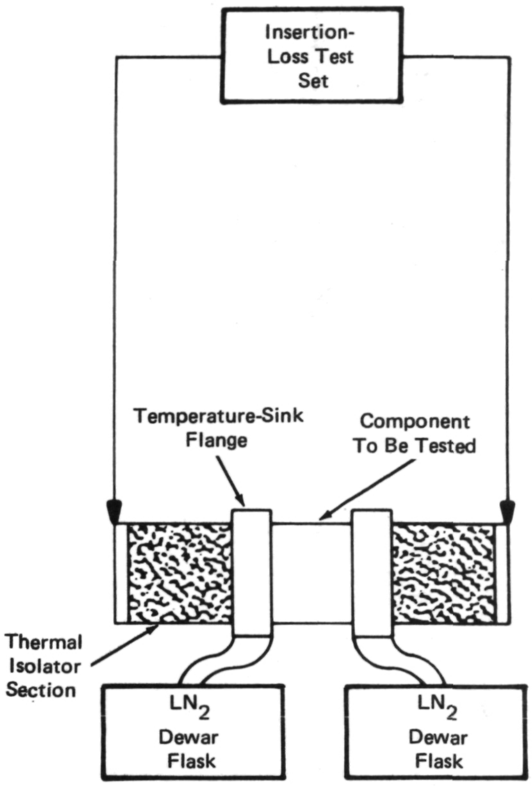
The idler tuner consists of the waveguide section, the fixed capacitive obstacle, and the screw-adjustable capacitive obstacle. These elements restrict the major part of the circulating idler current to the immediate vicinity of the diodes.

Each wafer has its own adjustable idler resonator; this is a key feature in the attainment of both low-noise operation and broad bandwidth. The diode is reactance matched at the signal frequency by coaxial contact assembly and an external matching structure, which effectively presents the negative of the average junction reactance to the junction point of the two diodes. This arrangement also efficiently couples energy to and from the diode without affecting the idler current.

Source: Lawrence E. Dickens of
Westinghouse Electric Corp.
under contract to
Goddard Space Flight Center
(GSC-11617)

Circle 18 on Reader Service Card.

TRANSMISSION-LINE LOSS MEASUREMENTS AT CONTROLLED TEMPERATURES



Insertion-Loss Test System

The figure illustrates an apparatus for measuring the insertion loss of transmission-line components, such as waveguides or coaxial lines, as a function of the temperature of the transmission line. With the new apparatus, the insertion loss of any transmission line or transmission-line component can be measured over a wide range of temperatures and frequencies.

The insertion-loss test set as shown is connected to thermal isolator sections which are connected in turn to temperature-sink flanges and to the component to be tested. The flanges incorporate cooling or heating means for establishing the required temperatures. In the case illustrated, the flanges are brazed to a pair of liquid-nitrogen dewar flasks. The insertion loss of the system, with the component in place, is measured directly with this configuration.

After testing, the component is removed, and the temperature-sink flanges are connected so that the isolator elements and the flanges are the only elements in the test path. Subtracting the second reading from the first reading gives the insertion loss of the component under test.

Source: C. T. Stelzried and
D. L. Mullen of
California Institute of Technology
under contract to
NASA Pasadena Office
(NPO-10717)

Circle 19 on Reader Service Card.

SIGNAL POWER MEASUREMENT: DETECTOR SUBSYSTEM

The radio-frequency-interference (RFI) measuring system consists of a predetection, band-limiting, interference filter; an interference detector; and a post-detection filter. Since it is estimated that by 1985 a major portion of international voice and data communications will be carried by Earth-orbiting communications satellites, the system may be of value in aiding the research and development of more efficient ground-to-satellite and satellite-to-ground broadcast and reception systems. Practical measurement and analysis of RFI to satellites from Earth sources has not been adequate. The system measures and identifies all interfering Earth radio transmitting sources radiating in the 5925-MHz to 6425-MHz band, according to power-density levels, carrier-frequency distribution, and geographical location.

In operation, the signal is switched on. After passing through the bandpass filter, the signal is rectified and filtered by an RC network. The output of the post-detection filter is fed to the processor which takes the required number of samples. The system then is reset, and the procedure is repeated for each 10-kHz segment.

The number of samples required for a prescribed accuracy is highly dependent on the randomness of the interfering signal. This number is bounded only by the characteristics of the CW signal and noise. The major aspects of an incoming signal are described by its mean value and the standard deviation. The total signal power is measured by the ratio of the deviation to the mean value. The significance of this ratio to the interfering qualities of the signal is under investigation.

Source: Samuel Sabaroff of
Hughes Aircraft Co.
under contract to
Goddard Space Flight Center
(GSC-11607)

Circle 20 on Reader Service Card.

Patent Information

The following innovations, described in this Compilation, have been patented or are being considered for patent action as indicated below:

Adaptive Detector FM/CW Radar (Page 2) MFS-22234

and

Hooded Antenna for Measurement of Electromagnetic Radiation in a Shielded Enclosure (Page 3) MFS-21240

and

Convoluted Audio Filter (Page 11) MFS-22729

Inquiries concerning rights for the commercial use of these inventions should be addressed to:

Patent Counsel
Marshall Space Flight Center
Code CC01
Marshall Space Flight Center, Alabama 35812

Compact Temperature Sensitive Telemetry Transmitter (Page 8) NPO-10649

This invention has been patented by NASA (U.S. Patent No. 3,541,450). Inquiries concerning nonexclusive or exclusive license for its commercial development should be addressed to:

Patent Counsel
NASA Pasadena Office
4800 Oak Grove Drive
Pasadena, California 91103

Rate Data Encoder (Page 16) LAR-10128

This invention has been patented by NASA (U.S. Patent No. 3,714,645). Inquiries concerning nonexclusive or exclusive license for its commercial development should be addressed to:

Patent Counsel
Langley Research Center
Mail Stop 313
Hampton, Virginia 23665

Pseudonoise Test Set for Real-Time and Non-Real-Time System Evaluation Using an RMS Error Criterion (Page 20) MFS-22671

This invention is owned by NASA, and a patent application has been filed. Inquiries concerning nonexclusive or exclusive license for its commercial development should be addressed to:

Patent Counsel
Marshall Space Flight Center
Code CC01
Marshall Space Flight Center, Alabama 35812

Varactor Diode Assembly (Page 29) GSC-11617

This invention has been patented by NASA (U.S. Patent No. 3,833,857). Inquiries concerning nonexclusive or exclusive license for its commercial development should be addressed to:

Patent Counsel
Goddard Space Flight Center
Code 204
Greenbelt, Maryland 20771

Notes:

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